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Proceedings of the IRE

in this issue

TIME-VARYING NETWORKS
COMPATIBLE SINGLE-SIDEBAND
CERAMIC MAGNETIC CIRCUITS
ELECTRON TRIGONOMETRY
FM-AM METHOD OF COMPRESSION
LOUDSPEAKER MEASUREMENT PRACTICES
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ABSTRACTS AND REFERENCES

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COMPATIBLE SINGLE-SIDEBAND: page 1503

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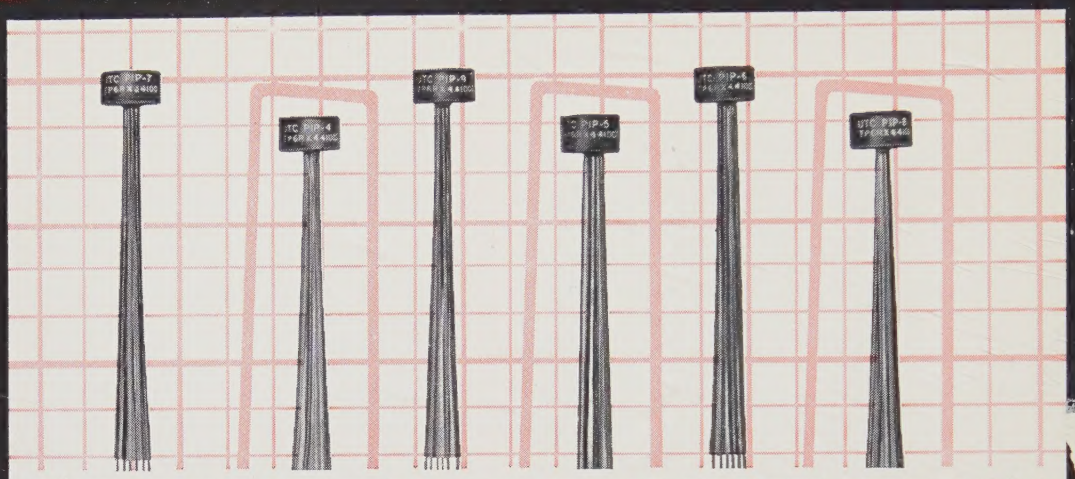

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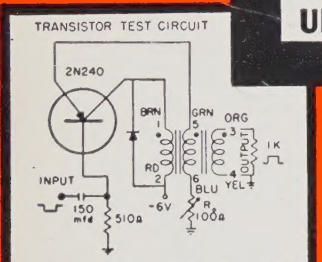

All units individually checked and adjusted, in transistor circuit illustrated, to parameters in table.

DEFINITIONS

Amplitude: Intersection of leading pulse edge with smooth curve approximating top of pulse.
Pulse width: Microseconds between 50% amplitude points on leading and trailing pulse edges.
Rise Time: Microseconds required to increase from 10% to 90% amplitude.
Overshoot: Percentage by which first excursion of pulse exceeds 100% amplitude.
Droop: Percentage reduction from 100% amplitude a specified time after 100% amplitude point.
Backswing: Negative swing after trailing edge as percentage of 100% amplitude.

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Type No.	1-Brn 2-Rd	3-Org 4-Yel	5-Grn 6-Blu	Width μ Sec.	Rise Time	% Over Shoot	Droop %	% Back Swing	P Width μ Sec.	Volt Out	Rise Time	% Over Shoot	Droop %	Back Swing	Imp. in, out,
PIP-1	.18	.20	.07	.05	.02	0	0	37	.05	9	.018	0	0	12	50
PIP-2	.47	.56	.17	.1	.025	0	0	25	.1	8	.02	0	0	5	50
PIP-3	1.01	1.25	.37	.2	.030	2	0	15	.2	7	.035	0	0	5	100
PIP-4	1.5	1.85	.54	.5	.05	0	0	15	.5	7	.06	0	0	0	100
PIP-5	2.45	3.1	.9	1	.08	0	0	14	1	6.8	.15	0	0	5	100
PIP-6	3.0	3.7	1.1	2	.10	0	0	15	2	6.6	.18	0	2	10	100
PIP-7	4.9	6.05	1.8	3	.20	0	0	14	3	6.8	.20	0	2	10	100
PIP-8	8.0	9.7	2.9	5	.30	0	0	3	5	7.9	.22	0	13	25	200
PIP-9	13.1	15.9	4.7	10	.35	0	5	12	10	6.5	.4	0	15	20	200
PIP-100	Transistor pulse transformer kit, consisting of PIP-1 thru PIP-9 in plastic case.														

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COVER

A single-sideband system that is compatible with AM was modulated by white noise to produce this panoramic picture. It shows how the spectrum is confined essentially within the lower sideband. The system is described on page 1503.

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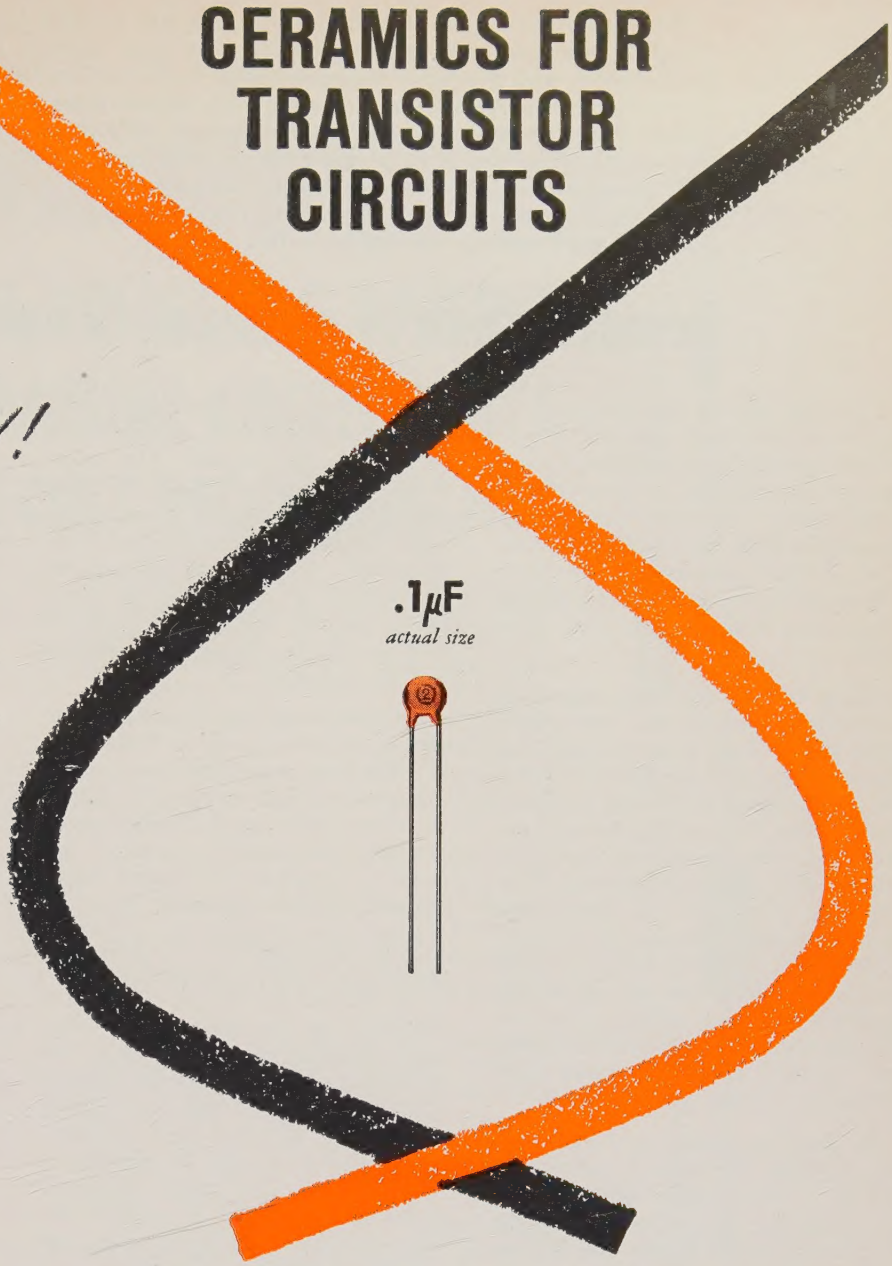

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2.2	.840	.47	.840	.1	.660

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Many new and unusual problems are brought to light by man's contemplated ventures into space. One of particular interest to us at AIL in connection with our space communication studies arises from the anticipated requirement for controlling a roving lunar vehicle from earth. Fred Hallden, one of our Consultants, designed a system with appropriate time delays to enable objective study of this problem with human "drivers." This month, Al Swanson, Jack Smith and Norm Freeberg of our Medical & Biological Physics Department, describe some interesting results of tests involving human performance capability in this situation.

Remote Control of Lunar Ground Vehicle

Once the general characteristics of the moon have been analyzed by low altitude orbiting lunar satellites, a more detailed investigation of lunar surface characteristics must be made before manned landings can be safely effected. This detailed investigation probably will be made by suitably instrumented lunar robot vehicles controlled from earth.

Many of the problems associated with reliable control of a lunar-based robot vehicle from an earth-based station are unique. Prime among these is the system delay introduced by signal transmission. The distance between the earth and moon sets the one-way signal transmission time at 1.32 seconds. Consequently, whenever an earth-based operator issues a command to the lunar vehicle there will be a minimum delay of 2.64 seconds before he receives any feedback from the vehicle as to how the command was received and carried out. Control problems, under such conditions, require study.

In order to obtain experience with the problems of remote manual control of a robot vehicle, when viewing an unfamiliar environment and with an unusual control-display time lag, we are currently conducting a series of simulation studies. These studies, employing a model vehicle, a mock-up lunar landscape, and a closed circuit video system, enable us to explore operator capabilities and determine problem areas which might not otherwise be apparent until much later in system development.

Our simulator vehicle is a small tractor, electrically powered chassis with treads, on which a forward looking television camera is mounted. The vehicle is controlled from a separate control room in which a video monitor is located. This displays a "windshield" view as seen by the camera mounted on the vehicle. Control of the vehicle is based almost entirely upon the information perceived by an operator using this display. Commands necessary to drive the vehicle are issued by the operator using twin control levers each of which controls one of the two vehicle tracks. The commands are converted into binary code, in which form they enter a shift register. The output of the shift register energizes relays which route appropriate power to the vehicle and serve to execute the desired commands. A delay is introduced into the system by controlling the clock rate of the shift register and hence the time between the issuing of a command and its execution.

The mock-up lunar landscape consists of an area of about 400 sq. ft. in which was constructed a course consisting of several irregularly shaped obstacles. When displayed on the monitor, the simulated "moonscape" appears as a group of bright objects against a dark background.

With this equipment we have conducted a series of experiments designed to investigate the effects of different conditions on an operator's performance. A group of eight subjects was required to drive the vehicle through a series of courses with and without a time delay and to reach a predetermined goal as quickly as possible by whatever route they chose. Their performance was graded in terms of distance traveled, total travel time, and number of obstacle collisions.

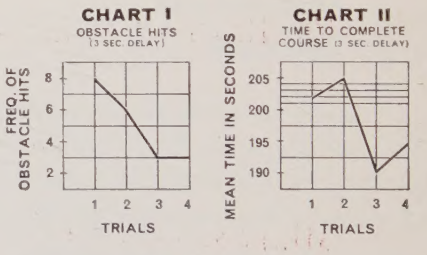
The results of these experiments permit us to draw certain conclusions and to define problems for further study. Among these are the following:

1. Geographical orientation is difficult to maintain by video means in an unfamiliar landscape with numerous obstacles. As a result, we are currently developing methods to display information which will enable the operator to maintain his orientation (e.g., navigation and radar displays).

2. The flat plane video presentation utilized here deprives the operator of certain valuable visual cues. Estimation of closure rate and size judgment are difficult. Therefore, we are developing a method for presenting a stereoscopic video display to the operator. This device will be incorporated and specific advantages of stereoscopic viewing determined.

3. It is not possible for an operator to run the vehicle continuously with any degree of accuracy under a 3 second command delay. Approximately two anticipatory control commands are possible, so that the operator must stop; visually search and then start again. Traversing a course took about twice as long with the delay as without it.

4. For the limited number of trials run under the 3 second delay, operators tended to become more proficient the longer they operated the vehicle (charts I and II). The limits of this proficiency have not yet been determined, but it is assumed that with proper training and appropriate displays an operator will be able to control, with reasonable accuracy, a vehicle having a substantial response time lag.



Our simulator will be improved upon and should enable us to learn a great deal more about the display-control problems faced by the operators of a remotely controlled vehicle. It will also serve as the test bed for the development of advanced display-control techniques.

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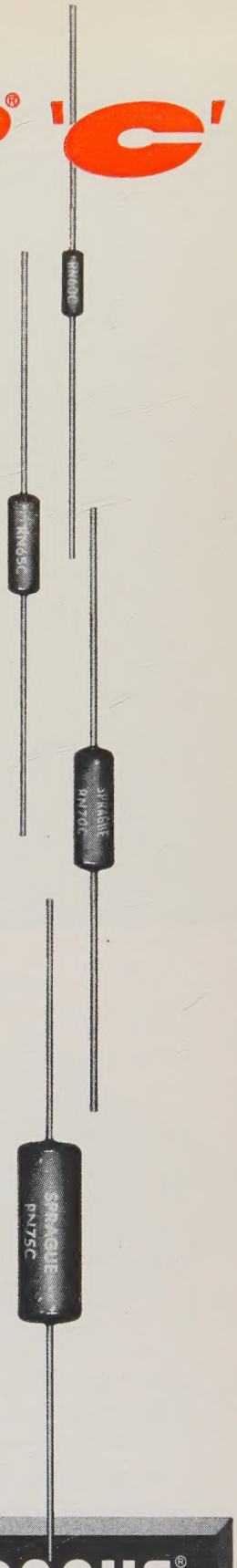

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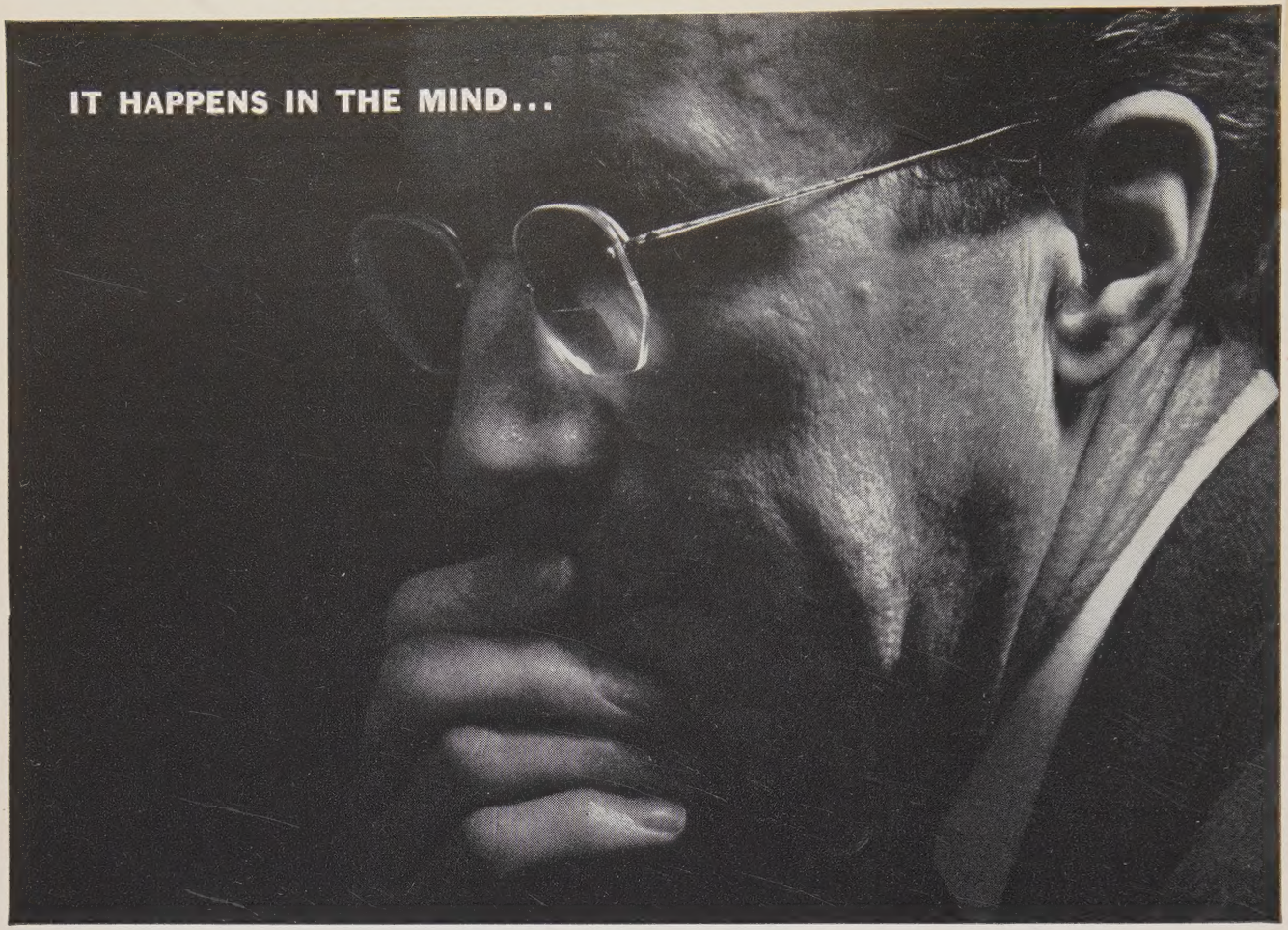

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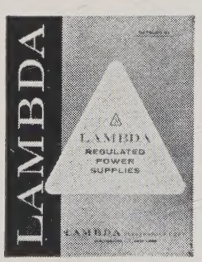

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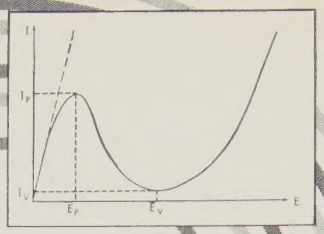

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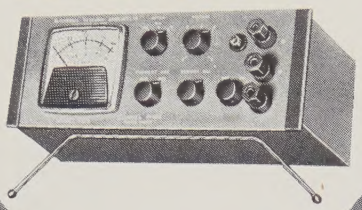

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- **Incorporation into equipment for supplying power at extremely low voltage levels for biasing tunnel diode oscillators, amplifiers, trigger and magnetic circuits.**

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Input	105-125VAC, 60-400cps	Line Regulation	within 0.05% for full input change
Output	0-6 VDC @ 0-100 ma	Load Regulation	within 0.05% or 2 mv for full load change
Vernier Adjustment	±0.5 volts	Temperature Coefficient	approx. 0.01%/°C
Internal Impedance	less than 0.01 ohms	Size	3½" x 7" x 9½"
Ripple	less than 1 mv rms		
Price (FOB FACTORY) Model TD6M—$275			

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Laboratories and Factory: 67 Factory Place, Cedar Grove, New Jersey CEnter 9-3000

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Meetings with Exhibits

● As a service both to Members and the industry, we will endeavor to record in this column each month those meetings of IRE, its sections and professional groups, which include exhibits.

Δ

October 9-11, 1961

National Electronics Conference, International Amphitheatre, Chicago, Ill.

Exhibits: Mr. Rudy Napolitan, National Electronics Conference, 228 N. LaSalle St., Chicago, Ill.

October 19-20, 1961

Symposium on Electronics, Engineering and Education, Greensboro Coliseum, Greensboro, N.C.

Exhibits: Mr. H. G. Eidson, Jr., Dept. 8760, Charham Road Plant, Western Electric Co., Inc., Winston-Salem, N.C.

October 23-25, 1961

East Coast Conference on Aerospace & Navigational Electronics, Lord Baltimore Hotel, Baltimore, Md.

Exhibits: Mr. Robert J. Henderson, Martin Company, Ground Support Equipment Dept., Baltimore, Md.

October 24-26, 1961

Eighth Annual Meeting, Professional Group on Nuclear Science Symposium on Aero-Space Nuclear Propulsion, Hotel Riviera, Las Vegas, Nevada

Exhibits: Mr. D. J. Niehaus, Bendix Research Labs. Div., Southfield, Mich.

October 26-27, 1961

Instrumentation Facilities for Biomedical Research Symposium, Sheraton Fontenelle Hotel, Omaha, Neb.

Exhibits: Mr. Harold G. Beenken, University of Nebraska, College of Medicine, 42 & Dewey Avenue, Omaha, Neb.

November 6-8, 1961

Special Technical Conference on Non-Linear Magnetics, Statler Hilton Hotel, Los Angeles, Calif.

Exhibits: Mr. Philip Diamond, Perkin Electronics Corp., 345 Kansas St., El Segundo, Calif.

November 13-16, 1961

Seventh Annual Conference on Magnetism & Magnetic Materials, Hotel Westward Ho, Phoenix, Ariz.

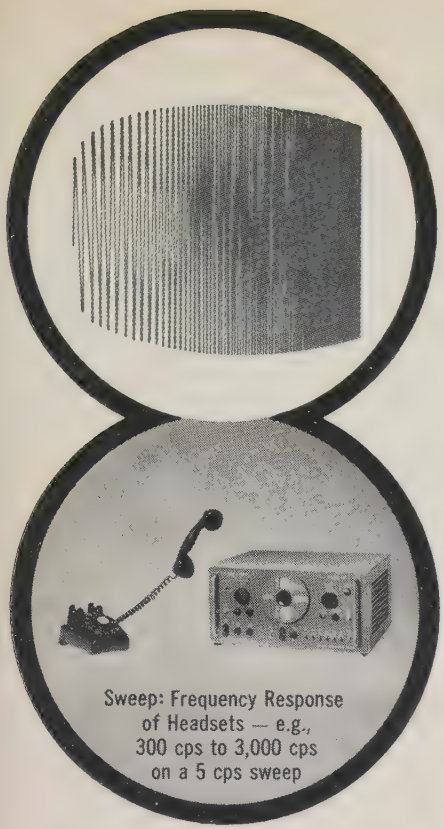
Exhibits: Mr. John L. Whitlock, 253 Waples Mill Road, Oakton, Va.

November 14-16, 1961

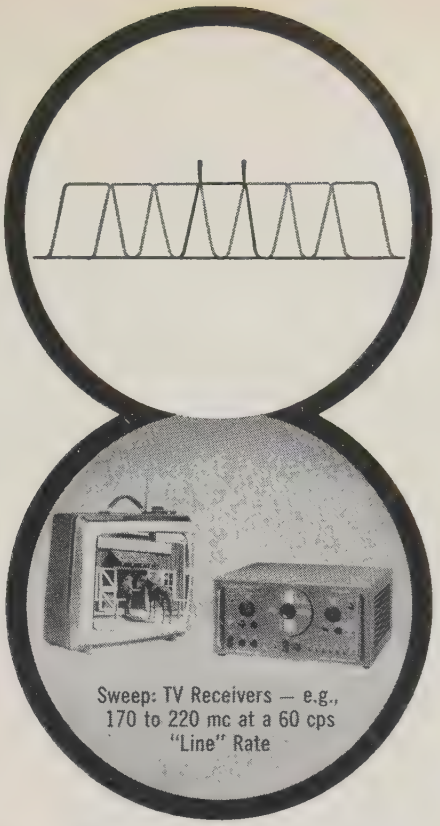
Northeast Research and Engineering Meeting (NEREM), Somerset Hotel & Commonwealth Armory, Boston, Mass.

Exhibits: Mr. Stewart K. Gibson, Instruments of New England, 108 Greenwood Lane, Waltham 54, Mass.

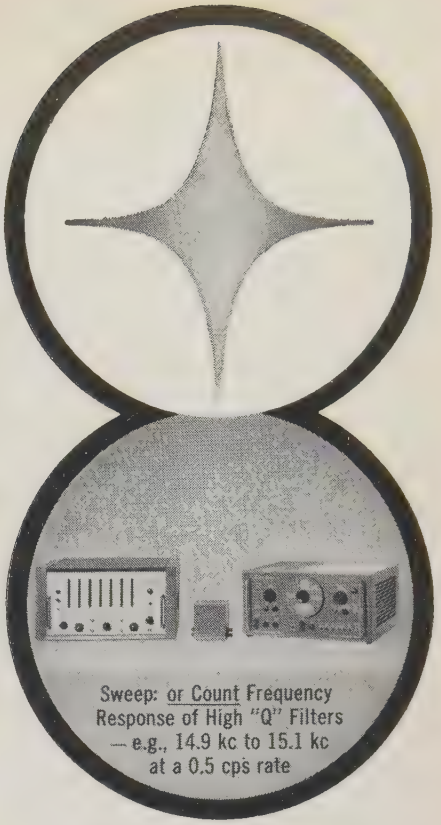
(Continued on page 10A)



Sweep: Frequency Response of Headsets — e.g., 300 cps to 3,000 cps on a 5 cps sweep



Sweep: TV Receivers — e.g., 170 to 220 mc at a 60 cps "Line" Rate



Sweep: or Count Frequency Response of High "Q" Filters — e.g., 14.9 kc to 15.1 kc at a 0.5 cps rate

Complete VERSATILITY... *Audio, Video, VHF*

KAY *Ligna-Sweep* SKV®

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SWEEPING OSCILLATOR

935-B

50 CPS TO 220 MC IN 12 BANDS • WIDE RANGE OF SWEEP WIDTHS VARIABLE REP RATES • MANUAL AND AUTOMATIC OPERATION

- **Single wide-sweep video displays from 10 mc down to 1 kc.**
- **Linear and logarithmic sweeps of 0.2 cps to 30 cps; or sweep locked to line frequency.**
- **Audio Sweep of 50 cps to 20,000 cps.**
- **8 fixed, narrow-band video frequency sweeps for repetitive operations.**
- **Fundamental frequency 10 mc to 220 mc (widths to 30 mc plus).**
- **High-level output of 1 V rms into 70 ohms. AGC'd to ± 0.5 db over widest sweep.**
- **Manually-operated control for varying oscillator frequency.**
- **Fixed pulse-type markers or variable marker provision.**

Price: $1295.00 F.O.B., Factory ($1425.00 F.A.S., New York).

The wide frequency range, extensive choice of sweep widths and repetition rates make the Kay *Ligna-Sweep* SKV a most useful sweeping oscillator.

For high frequency work, the unit provides 9 sweep bands, operating at fundamental frequencies for wide, stable sweeps from 10 to 220 mc. At the low end of the spectrum, an audio frequency sweep from 50 to 20,000 cps is provided. High order stability permits frequency sweeps to as low as 50 cps.

For checking high-Q circuits and low-frequency response characteristics, either log or linear sweeps at variable rep rates down to 0.2 cps are available. This wide choice of sweep rates (continuous to 30 cycles, and fixed line lock) makes it easy to select that highest *rep rate* which gives both an *accurate response* display and easiest, brightest viewing on the scope screen. With the manual frequency control, the trace on the scope screen may be held and examined in detail, (counted precisely, measured on a VTVM) at any frequency point on the scope display.

Check Video Bandpass — e.g., 1 kc to 10 mc at 1 cps rate

Check Radar IF's — e.g., 25 to 35 mc at 30 cps

Check Audio Bandpass — e.g., 50 cps to 20 kc at 1 cps Log Sweep

Write for Complete Catalog Information

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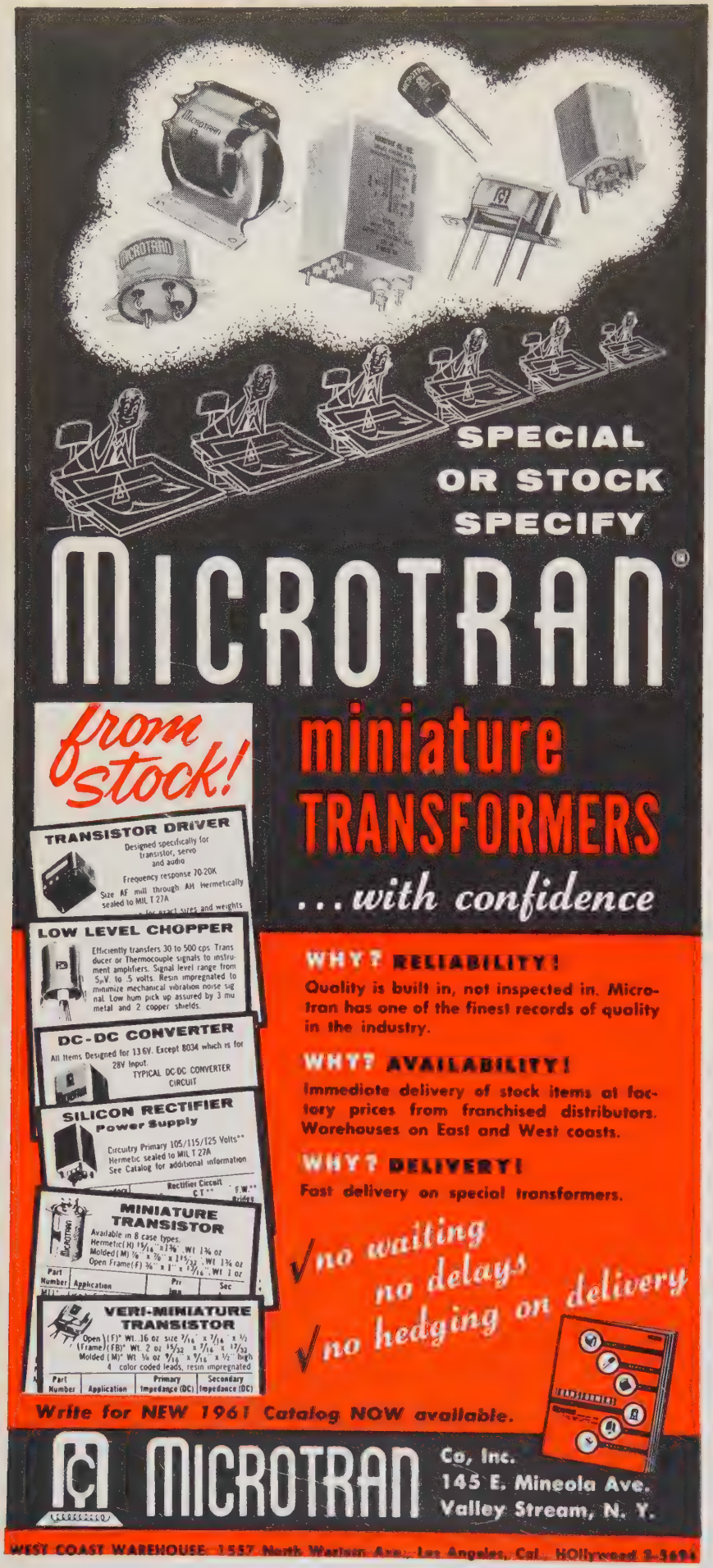
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Designed specifically for transistor, servo and audio
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Efficiently transfers 30 to 500 cps Transducer or Thermocouple signals to instrument amplifiers. Signal level range from .5μV. to .5 volts. Resin impregnated to minimize mechanical vibration noise signal. Low hum pick up assured by 3 mu metal and 2 copper shields.
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All Items Designed for 13.6V. Except 8034 which is for 28V Input.
TYPICAL DC-DC CONVERTER CIRCUIT
SILICON RECTIFIER
Power Supply
Circuitry Primary 105/115/125 Volts**
Hermetic sealed to MIL T 27A
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MINIATURE TRANSISTOR
Available in 8 case types.
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Application
VERI-MINIATURE TRANSISTOR
Part Number
Application
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Quality is built in, not inspected in. Microtran has one of the finest records of quality in the industry.
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Meetings with Exhibits

(Continued from page 8A)

November 30-December 1, 1961

Professional Group on Vehicular Communications Conference, Hotel Radisson, Minneapolis, Minn.

Exhibits: Mr. Jean Poole, General Electric Co., 3280 Gorham Ave., Minneapolis, Minn.

December 12-14, 1961

Eastern Joint Computer Conference, Sheraton-Park Hotel, Washington, D.C.

Exhibits: Mr. Charles Phillips, 5603 Jordan Road, Washington 16, D.C.

February 7-9, 1962

3rd Winter Convention on Military Electronics, Ambassador Hotel, Los Angeles, Calif.

Exhibits: Mr. Walter E. Peterson, Automation Development Corp., 11824 Jefferson Blvd., Culver City, Calif.

March 1-3, 1962

Eighth Scintillation and Semiconductor Counter Symposium, Shoreham Hotel, Washington, D.C.

Exhibits: Dr. George A. Morton, RCA Labs., Princeton, N.J.

March 26-29, 1962

International Radio & Electronics Show and IRE International Convention, Waldorf-Astoria Hotel and New York Coliseum, New York, N.Y.

Exhibits: Mr. William C. Copp, IRE Advertising Dept., 72 West 45th St., New York 36, N.Y.

April 11-13, 1962

SWIRECO (South West IRE Conference & Electronics Show), Rice Hotel, Houston, Texas

Exhibits: Mr. R. J. Loofbourrow, Texaco Company, P.O. Box 425, Bellaire, Texas

May 1-3, 1962

Spring Joint Computer Conference, Fairmont Hotel, San Francisco, Calif.

Exhibits: Mr. John W. Ball, Pacific Telephone Co., 3240 Arden Way, Sacramento, Calif.

May 14-16, 1962

NAECON (National Aerospace Electronics Conference), Miami & Biltmore Hotels, Dayton, Ohio

Exhibits: IRE Dayton Office, 33 N. Torrence St., Dayton, Ohio

May 23-25, 1962

National Telemetering Conference, Sheraton Park Hotel, Washington, D.C.

Exhibits: Mr. H. W. Royce, Martin Company, Baltimore, Md.

Δ

Note on Professional Group Meetings: Some of the Professional Groups conduct meetings at which there are exhibits. Working committeemen on these groups are asked to send advance data to this column for publicity information. You may address these notices to the Advertising Department and of course listings are free to IRE Professional Groups.

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2NI658 2NI659

Looking for a compact 1 to 10 watt Germanium Power Transistor with this kind of top level performance?

ACTUAL SIZE

Top reliability is built into each Honeywell transistor. Vacuum baked, given rigorous life and dynamic tests, each is enclosed in a new cold-welded miniature case with threaded mounting stud.

Designed for general use, the Honeywell 2N1658 and 2N1659 are ideal for servo amplifiers, relay drivers, high power unit driver stages, and as medium power switches. Measuring only 0.35 inches in diameter, they handle one to ten watts!

You'll want data sheets, including characteristic curves and servo amplifier circuit diagram, for your files. Fill out and mail the coupon today!

- *Leakage*—I_{CBO} less than 0.5 ma at 80v, 2N1658—60v, 2N1659
- *Saturation voltage*—$V_{CE\ (SAT)}$ less than 0.25v
- *Floating potential*—V_{EBF} less than 0.25v
- *Thermal resistance*—$\Theta_{J\text{-}MB}$ 5° C/watt max. (T_J max., 100° C)
- *Gain*—h_{FE} 30 min @ 0.2 amp; h_{FE} 20 min @ 1.0 amp
- *Frequency response*—500 kc minimum gain bandwidth product
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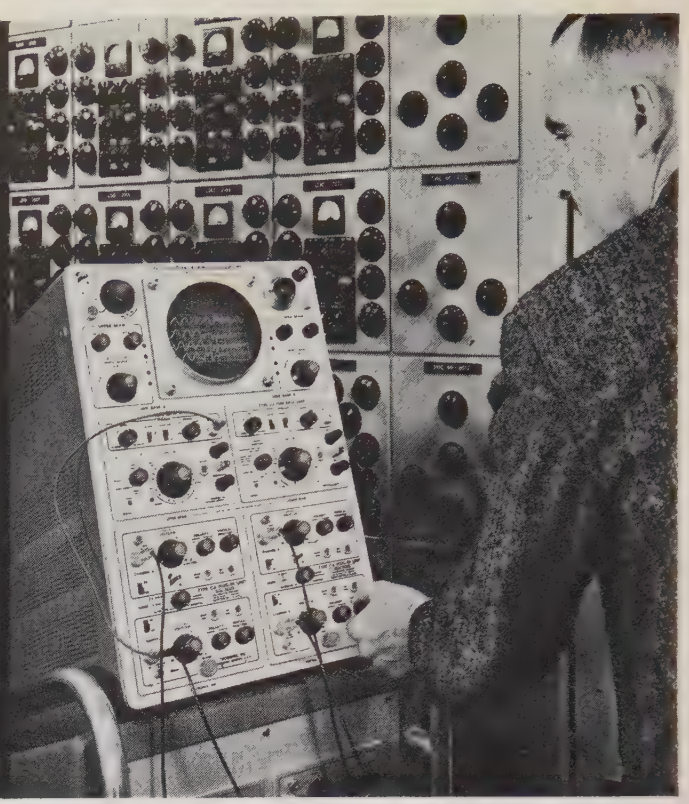
DUAL-BEAM OSCILLOSCOPE

with Sweep Delay

Independent X and Y Deflection

DC-to-30 MC, 12-nanosecond risetime with fast-rise plug-in units



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CHARACTERISTICS

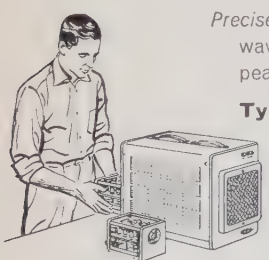
Adaptable Vertical System—accepts interchangeable plug-in preamplifiers.

Versatile Sweep Features—wide range from 0.1 μsec/cm to 5 sec/cm in 24 calibrated main sweep rates, continuously variable uncalibrated to 12 sec/cm. 5X magnifier increases calibrated sweep time to 20 nsec/cm. Single sweep facilitates recording one-shot phenomena.

Calibrated Sweep Delay—two modes of jitter-free operation.

Complete Triggering Facilities—amplitude-level (manual) selection or fully automatic control.

High Writing Rate—10-KV accelerating potential provides bright traces at low repetition rates. 4 by 10 centimeter display for each beam, with 2 centimeter overlap.



Precise Amplitude Calibrator—with 18 square-wave voltages (from 0.2 mv to 100 v peak-to-peak) available at the front panel.

Type 555 (without preamplifiers) . . $2600
Includes Indicator Unit, Power Unit, 2 Time-Base Units, 4 Probes, Time-Base Extension, 7 other accessories.

Price f.o.b. factory

For a demonstration of the capabilities of a Type 555 in your own waveform-comparison analyses, please call your Tektronix Field Engineer.

At Bonneville Power Administration—in one of many continuing studies with their transient analyzer—a System Engineer uses a Tektronix Type 555 Oscilloscope to display fast switching transients occurring in each phase of a transmission system. By observing the three switching-transient traces with a timing trace, he can quickly and easily compare magnitudes, phase relationships, transient times, and accurately determine overvoltage characteristics of the system.

Although invaluable in displaying up to four traces at once—with Tektronix dual-trace units in both channels—the Type 555 adapts easily to almost every oscilloscope application in the dc-to-30 mc range.

For example, you can control either or both beams with either time-base generator. You can operate one time-base unit as a delay generator—*hold off the start of any sweep generated by the other* for a precise interval from one-half microsecond to 50 seconds—and observe both the original display and the delayed display at the same time. *You can interchange any combination of 17 "letter-series" plug-in units* for signal-handling ease and versatility in waveform-comparison analyses, such as dual-beam pulse-sampling, transistor-rise-time testing, semiconductor-diode-recovery-time studies, strain gage and other transducer measurements, differential-comparator applications, as well as multiple-trace work in general laboratory experiments.

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ENGINEERING REPRESENTATIVES: Kentron Hawaii Ltd., Honolulu, Hawaii. Tektronix is represented in twenty-five overseas countries by qualified engineering organizations.

European and African countries, the countries of Lebanon and Turkey, please contact TEKTRONIX INTERNATIONAL A.G., Terrassenweg 1A, Zug, Switzerland, for the name of your local engineering representative. Other Overseas areas, please write or cable directly to Tektronix, Inc., International Marketing Department, P. O. Box 500, Beaverton, Oregon, U.S.A. Cable: TEKTRONIX.

C TYPE MAGNETS in a wide range of sizes to meet your design needs in ★Transverse Field Isolators ★Differential Phase Shifters ★Duplexers

Arnold C-type Alnico Magnets are available in a wide selection of gap densities ranging from 1,000 to over 7,500 gausses. There are six different basic configurations with a wide range of stock sizes in each group.

The over-all size and gap density requirements of many prototype designs can be met with stock sizes of Arnold C Magnets, or readily supplied in production quantities.

When used in transverse field isolators, Arnold C Magnets supply the magnetizing field to bias the ferrite into the region of resonance, thus preventing interaction between microwave networks and isolating the receiver from the transmitter. These magnets are also used in differential phase shifters and duplexers, and Arnold is prepared to design and supply tubular magnets to provide axial fields in circular wave guides.

A feature of all Arnold C Magnets is the excellent field uniformity along the length of the magnet. Versatility in design may be realized by using multiple lengths of the same size magnet stacked to accomplish the needs of your magnetic structure.

Let us work with you on any requirement for permanent magnets, tape cores or powder cores. ● For information on Arnold C Magnets, write for Bulletin PM-115. Address *The Arnold Engineering Company, Marengo, Illinois.*

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IRE News and Radio Notes

Current IRE Statistics

(As of August 31, 1961)

Membership—90,557
Sections*—110
Subsections*—31
Professional Groups*—28
Professional Group Chapters—287
Student Branches†—211

* See this issue for a list.
† See June, 1961, issue for a list.

Calendar of Coming Events and Authors' Deadlines*

1961

Oct. 1–6: CISPR, Univ. of Pennsylvania, Philadelphia.

Oct. 2–3: 29th Ann. Mtg. Engineers' Council for Professional Dev., Sheraton Seelbach Hotel, Louisville, Ky.

Oct. 2–4: 7th Nat'l. Communications Symp., Utica, N. Y.

Oct. 2–4: IRE Canadian Electronics Conf., Automotive Bldg., Exhibition Park, Toronto, Canada.

Oct. 6–7: 11th Ann. Broadcast Symp., Willard Hotel, Washington, D. C.

Oct. 9–11: Nat'l. Electronics Conf., Int'l. Amphithea re, Chicago, Ill.

Oct. 9–15: ARS Space Flight Rept. to the Nation, New York Coliseum, New York, N. Y

Oct. 16–17: 5th Nat'l. Symp. on Engrg. Writing and Speech, Kellogg Ctr. for Continuing Education, Michigan State Univ., East Lansing.

Oct. 16–17: Int'l. Conf. on Ionization of the Air, Franklin Inst., Philadelphia, Pa.

Oct. 19–20: 6th Ann. North Carolina Section Symp., Greensboro Coliseum, Greensboro, N. C.

Oct. 23–25: East Coast Conf. on Aerospace & Navigational Electronics, Lord Baltimore Hotel, Baltimore, Md.

Oct. 23–25: URSI-IRE Fall Mtg., Univ. of Texas, Austin.

Oct. 23–26: PGNS 8th Ann. Mtg., Hotel Riviera, Las Vegas, Nev.

Oct. 26–27: Symp. on Instrumentation Facilities for Biomedical Res., Sheraton Fontenelle Hotel, Omaha, Neb.

Oct. 26–28: 1961 Electron Devices Mtg., Sheraton-Park Hotel, Washington, D. C.

Oct. 30–31: Radio Fall Mtg., Hotel Syracuse, Syracuse, N. Y.

Nov. 6–8: 6th Ann. Special Technical Conf. on Nonlinear Magnetics, Statler-Hilton Hotel, Los Angeles, Calif.

Nov. 7–9: 7th Conf. on Radio Interference Reduction and Electronic Compatibility, Ill. Inst. Tech., Chicago, Ill.

* DL = Deadline for submitting abstracts.

(*Continued on page 15A*)

Broadened Scope for N.B.S. Radio Propagation Journal

Since July, 1959, the *Journal of Research of the National Bureau of Standards* has appeared in four sections, each devoted to a main field of Bureau research and development. Of special interest to IRE members is Section D, which is edited by the Central Radio Propagation Laboratory. It is designed to provide a bimonthly publication whose subject matter concentrates on research in radio propagation, communications, and atmospheric physics. Topics covered include propagation in ionized media, scattering by turbulence, the effect of irregular terrain on propagation, diffraction and scattering by solid obstacles, propagation through time-varying media, and antennas. In the future, it is hoped that atmospheric physics will receive increased emphasis in the journal. The policy of inviting papers from outside N.B.S. will concontinue.

Dr. David M. Gates, who is Consultant to the Director of the Boulder Laboratories of N.B.S., has been the editor since July 1961. Associate editors are Thomas N. Gautier, N.B.S.; Dr. John P. Hagen, National Aeronautics and Space Administration, Washington, D.C.; Dr. Bernhard Haurwitz, High Altitude Observatory, Boulder, Colo.; Dr. Robert A. Helliwell, Radioscience Laboratory, Stanford University, Calif.; Richard C. Kirby, N.B.S.; Dr. C. Gordon Little, N.B.S.; Alvin G. McNish, N.B.S.; Kenneth A. Norton, N.B.S.; Dr. Franklin E. Roach, N.B.S.; Dr. Albert D. Wheelon, Space Technology Laboratories, Inc., Los Angeles, Calif.; and Dr. James R. Wait, N.B.S. (formerly the editor). IRE advisors are Dr. A. H. Waynick of Pennsylvania State University and Dr. E. K. Smith of N.B.S.

The full title of the journal is *Radio Propagation, Section D of the N.B.S., Journal of Research.*

The journal may be ordered from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C. The subscription price is $4.00 a year ($4.75 foreign). Make all checks payable to Superintendent of Documents.

Further information concerning the journal may be obtained from Mrs. Charlotte S. Landay, Assistant to the Editor, *Radio Propagation*, National Bureau of Standards, Boulder, Colo.

Plasma Sheath Symposium To Meet in April

The Second Symposium on The Plasma Sheath—Its Effect Upon Re-entry Communication and Detection, sponsored by the Air Force Cambridge Research Laboratories, L. G. Hanscom Field, Bedford, Mass., will be held April 10–12, 1962 at New England Mutual Hall, Boston, Mass. The purpose of this symposium is to exchange both classified and unclassified information on the problems of the transmission and reflection of electromagnetic radiation through the plasma sheath which surrounds a hypersonic vehicle, and their effect upon re-entry telemetry, communications, and radar detectability. Specific topics include:

The Effect of the Plasma Sheath Upon Communications and Telemetry Systems of Re-entry Vehicles.

Techniques for Improving Re-entry Communications Systems.

The Effect of the Plasma Sheath Upon Radar Cross-Section and Detectability.

Voltage Breakdown of Antennas at High Altitudes.

Flight Test Programs and Instrumentation for Re-entry Communication Tests.

Both classified and unclassified papers are solicited on the above and related topics. An abstract of approximately 200 words should be submitted to the address below prior to October 15, 1961. The classified abstracts should include an unclassified title for program use. Instructions for preparation of papers will be sent to each author.

Abstracts and requests for registration material should be submitted to Miss Alice Cahill, Code CRRD, Symposium Secretary, Air Force Cambridge Research Labs., L. G. Hanscom Field, Bedford, Mass.

Canadian Reliability Course Sponsored by PGQRC and ASQC

A Canadian Reliability Training Course will be held from October 30 to November 4, 1961, at the Sheraton Mount Royal Hotel in Montreal, Quebec. The schedule includes 36 hours of formal lectures supplemented by visual aids, discussions and guest speakers well-known in various fields of reliability. The syllabus presents detailed tuition in the management, theory and application of the most recent reliability techniques with the object of providing an over-all working knowledge of the tools and concepts required for the successful execution of the reliability program.

The course is limited to 35 people to insure close tutoring between student and instructor. This limited enrolment makes prompt application imperative.

The instructors are Harmon S. Bayer of Detroit, Mich. and Thomas A. Budne of Great Neck, N. Y. who have instructed this course on twelve previous occasions and who have wide experience in military and industrial reliability problems.

The course fee of $225.00 includes all materials and text books, all noon meals and one evening meal. Applications for registration and checks should be made payable to "Reliability Training Conference" and addressed to Mrs. Naomi McAfee, Apt. A8, 2106 Tucker Lane, Baltimore, Md.

The Canadian Host Committee includes L. C. Thomas, Chairman; M. A. Fraser, I. Kirkpatrick, C. I. Soucy, A. B. Oxley, H. Reiche and J. B. Fogarty. Further information about the course may be obtained from L. C. Thomas, Canadian Aviation Electronics Ltd., P.O. Box 2030, Montreal 9, Quebec.

Call for Papers

1962 IRE International Convention

March 26–29, 1962

Waldorf-Astoria Hotel and the New York Coliseum New York N. Y.

Prospective authors are requested to submit all of the following information by:

October 20, 1961

1. 100-word abstract *in triplicate*, title of paper, name and address
2. 500-word summary *in triplicate*, title of paper, name and address
3. Indicate the technical field in which your paper falls:

Aerospace & Navigation Electronics	Engineering Writing & Speech
Antennas & Propagation	Human Factors in Electronics
Audio	Industrial Electronics
Automatic Control	Information Theory
Bio-Medical Electronics	Instrumentation
Broadcast & Television Receivers	Microwave Theory & Techniques
Broadcasting	Military Electronics
Circuit Theory	Nuclear Science
Communications Systems	Product Engineering & Production
Component Parts	Radio Frequency Interference
Education	Reliability & Quality Control
Electron Devices	Space Electronics & Telemetry
Electronic Computers	Ultrasonic Engineering
Engineering Management	Vehicular Communications

Note: Only original papers, not published or presented prior to the 1962 IRE International Convention, will be considered. *Any necessary military or company clearance of papers must be granted prior to submission.*

Address all material to: Dr. Donald B. Sinclair, Chairman
1962 Technical Program Committee
The Institute of Radio Engineers, Inc.
1 East 79 Street, New York 21, N. Y.

Schedule of Sessions For ECCANE Announced

The schedule of sessions and keynote speaker for the Eighth East Coast Conference on Aerospace and Navigational Electronics has been announced. The Conference will be held at the Lord Baltimore Hotel, Baltimore, Md., October 23–25, 1961. The keynote speaker will be General Bernard A. Schriever, Commanding General of the Air Force System Command; he will speak at the banquet on Tuesday evening, October 24.

Several hundred aerospace and navigational scientists are expected to attend the meetings which will feature the presentation of 48 papers by 95 authors. As in past years, awards will be made to the two outstanding papers presented at the conference.

The conference is set up so that there will be four sessions each day, two morning sessions running concurrently from 9:30 A.M. until 12:30 P.M. and two afternoon sessions running concurrently from 2:00 P.M. until 5:00 P.M. The program is basically designed to offer scientists in this comparatively infant field an opportunity to hear about and discuss the most advanced ideas in this critical area. The conference, although only eight years old, is considered venerable in this still emerging field.

This year's Conference is headed by R. J. Henderson of The Martin Company. Assisting Mr. Henderson with plans for the conference are E. Thomas, Co-Chairman; W. Vergara, Papers Chairman; J. Thompson, Local Arrangement; J. Clark, Security; Dr. Carroll Barrack, Finance; S. Ellen, Special Adviser and J. Laverty, Publicity and Printing.

The sessions are scheduled to run as follows:

Monday Morning, October 23

Session 1.1—Antennas and Propagation
Session 1.2—Air Traffic and Navigation

Monday Afternoon

Session 2.1—Electronic Systems and Techniques
Session 2.2—Aerospace Communications AJ Techniques (Secret)

Tuesday Morning, October 24

Session 3.1—Microwave Components and Microminiaturization
Session 3.2—Aerospace Command and Control (Secret)

Tuesday Afternoon

Session 4.1—Satellite Navigation
Session 4.2—Advanced Aerospace Electronics (Secret)

Wednesday Morning, October 25

Session 5.1—Satellite Communication and Control
Session 5.2—Real Time Aerospace Data Processing

Wednesday Afternoon

Session 6.1—Radar
Session 6.2—Computers and Data Processing

Calendar of Coming Events and Authors' Deadlines*

(*Continued from page 14A*)

Nov. 13–14: Conf. on Electrically-Exploded Wires, Kenmore Hotel, Boston, Mass.

Nov. 13–16: 7th Ann. Conf. on Magnetism and Magnetic Materials, Hotel Westward Ho, Phoenix, Ariz.

Nov. 14–16: NEREM, Boston, Mass.

Nov. 14: Electronic Systems Reliability Symp., Linda Hall Library Auditorium, Kansas City, Mo.

Nov. 30–Dec. 1: PGVC Conf., Hotel Radisson, Minneapolis, Minn.

Dec. 12–14: Eastern Joint Computer Conf., Sheraton-Park Hotel, Washington, D. C.

1962

Jan. 9–11: 8th Nat'l. Symp. on Reliability and Quality Control, Statler Hilton Hotel, Washington, D. C.

Jan. 28–Feb. 2: AIEE 1962 Winter General Mtg., New York, N. Y.

Feb. 7–9: 3rd Winter Conv. on Military Electronics, Ambassador Hotel, Los Angeles, Calif. (DL*: Nov. 15, 1961, M. E. Brady, Space Tech. Labs., P.O. Box 95001, Los Angeles 45, Calif.)

Feb. 14–16: Internat'l. Solid State Circuits Conf., Philadelphia, Pa. (DL*: Nov. 1, 1961, R. B. Adler, Lincoln Lab., MIT, Lexington, Mass.)

Mar. 1–3: 8th Scintillation and Semiconductor Counter Symp., Shoreham Hotel, Washington, D. C.

Mar. 26–29: Internat'l. Conv., Coliseum and Waldorf Astoria Hotel, New York, N. Y. (DL*: Oct. 20, 1961, D. B. Sinclair, IRE, 1 E. 79 St., New York, N. Y.)

Apr. 10–12: 2nd Symp. on the Plasma Sheath, England Mutual Hall, Boston, Mass. (DL*: Oct. 15, 1961, A. Cahill, Air Force Cambridge Res. Labs., Hanscom Field, Bedford, Mass.)

Apr. 11–13: SWIRE Conf. and Electronics Show, Rice Hotel, Houston, Tex. (DL*: Oct. 1, 1961, M. Graham, Rice Univ. Computer Project, Houston 1, Tex.)

May 1–3: Spring Joint Computer Conference, Fairmont Hotel, San Francisco, Calif. (DL*: Nov. 10, 1961, R. I. Tanaka, Lockheed Missiles and Space Co., Palo Alto, Calif.)

May 8–10: Electronic Components Conf., Marriott Twin Bridges Hotel, Washington, D. C. (DL*: Oct. 9, 1961, H. A. Stone, Bell Telephone Labs., Murray Hill, N. J.)

May 14–16: NAECON, Dayton, Ohio.

May 22–24: Natl. Microwave Theory and Techniques Symp., Boulder, Colo. (DL*: Dec. 18, 1961, R. W. Beatty, Natl. Bur. Standards, Boulder, Colo.)

May 23–25: Natl. Telemetering Conf., Sheraton Park Hotel, Washington, D. C.

June 25–27: 6th Natl. Convention on Military Electronics, Shoreham Hotel, Washington, D. C.

* DL = Deadline for submitting abstracts.

Symposium Program Announced For North Carolina Section

The Sixth Annual Symposium of the North Carolina Section, IRE, which will be held October 19–20, 1961 at the Coliseum, Greensboro, N. C., has for its theme "Electronics, Engineering and Education." The tentative program has been announced as follows:

Thursday Morning, October 19
Registration and Exhibits

Thursday Afternoon

Chairman: *W. A. Bernath, Western Electric Co.*

Welcoming remarks and introduction of the keynote speaker, *W. A. Bernath, Western Electric Co.*

Keynote Address, (to be announced)

Introduction of the next speaker by the Program Chairman, *G. W. Edwards, Western Electric Co.*

"Satellite Communications," *H. S. Black Bell Telephone Labs.*

"The Revolution in Engineering Education," *C. R. Vail, Dept. of Elec. Engrg., Duke University, Durham, N. C.*

(Speaker to be announced)

Thursday Evening
Speakers Banquet

"New Developments at Goodyear," *H. L. Flowers, Dir. Weapons Systems, Goodyear Rubber Co.*

(Speaker to be announced)

Friday Morning, October 20
Registration and Exhibits

Friday Afternoon

"Compatible Single-Sideband and the Kahn AM Stereophonic System," *L. P. Kahn, Kahn Research Labs., Inc., Freeport, L. I., N. Y.*

"Technicians in Industry," *W. R. Halstead, Gaston Technical Inst., Gastonia, N. C.*

(Speaker to be announced)

"The Maser," *T. E. Mardis, Bell Telephone Labs.*

Friday Evening
Banquet

Message from the Section Chairman: *A. B. MacIntyre, Dir. of Engrg., WUNC (TV), Chapel Hill, N. C.*

Banquet Address: *J. F. Byrne, Systems Research Lab., Motorola Inc., Riverside, Calif.*

Closing remarks: *W. A. Bernath*

MAECON Symposium Program Announced

The 1961 MAECON Symposium sponsored by the Kansas City Section of the IRE will be held at the Hotel Phillips in Kansas City on November 14, 1961. This year's meeting will emphasize the design and manufacturing aspects of the "Reliability of Electronic Systems."

Registration fee for the symposium (including luncheon) is $10.00. All registrants will receive preliminary copies of the papers several days prior to the meeting, and will be expected to become familiar with their contents. The authors will give only a brief resume during the meeting, with a major portion of the allotted period being devoted to discussion and questions. A final copy of the proceedings, including both papers and discussion, will be provided to each registrant at no additional cost.

For further information, or pre-registration, write to: F. A. Spies, The Bendix Corp., P.O. 1159, Kansas City 41, Mo.

Tuesday Morning, November 14

Moderator: *R. M. Somers, Director of Engineering, Bendix Corp., Kansas City, Mo.*

"Some Organizational Inferences Resulting from AGREE Reliability Testing," *G. W. Lindsay, Staff Aeronautical Systems Div., Wright-Patterson Air Force Base.*

"Achievement of Reliability in Electronic Parts," *H. T. Gruber, Reliability Engineering Div., Batelle Memorial Inst.*

"Meeting AGREE Reliability Requirements for Airborne TACAN Equipment, History and Results," *A. L. Floyd, Director of Plans and Programs, Hoffman Electronics Corp., Los Angeles, Calif.*

Tuesday Luncheon

Speaker: *Major General John B. Medaris (USA ret.), President, The Lionel Corp., Irvington, N. J.*

Toastmaster: *Dr. Charles N. Kimball, President, Midwest Research Inst., Kansas City, Mo.*

Tuesday Afternoon

Moderator: *L. D. Lanford, Engineering Superintendent, Western Electric Co.*

"The Second Reliability Problem—The Development-Manufacturing Interface," *A. J. Finocchi, Dir., Reliability Engineering, ITT Labs., Nutley, N. J.*

"What Have You Proved?" *W. E. Boyes, Supervisor, Reliability Dept., Sandia Corp., Albuquerque, N. Mex.*

PGMIL Calls for Papers for 1962 Winter Convention

A call for papers has been issued for the 1962 National Winter Convention on Military Electronics by its General Chairman, M. D. McFarlane. The Convention, which will be held at the Ambassador Hotel in Los Angeles, February 7–9, 1962, will be keynoted by a panel of assistant secretaries of the armed services (Research and Development). The panel will be moderated by the Honorable John H. Rubel, Assistant Secretary of Defense, Deputy Director of Defense Research and Engineering.

Unclassified and confidential classified papers will be presented in the various technical sessions. Confidential papers will cover: Undersea Warfare and Sonar Systems; Radar and Fire Control Systems; Aerospace Ground Equipment; Information and Data Handling Systems; Reconnaissance and Electronic Warfare; and, Military Systems Requirements and Environments. Unclassified papers will cover: System and Technical Management; Instrumentation; Reliability; Navigation and Air Traffic Control Systems; Military Equipment Design and Product Engineering. Both unclassified and classified sessions will be held on two subjects: Tracking, Telemetry and Command Systems; and Missile and Space Systems and Subsystems. The confidential classified sessions will be sponsored by the U. S. Air Force Systems Command.

Prospective authors are requested to submit to the Program Committee, prior to November 15th, an unclassified 100 word abstract, a 500 word summary, and a short biography. This information should be sent to Matthew E. Brady, Space Technology Laboratories, P.O. Box 95001, Los Angeles 45, Calif. Confidential material should be submitted to Maj. James L. Blilie, USAF, U. S. Air Force Systems Command, Regional Office, 6331 Hollywood Blvd., Los Angeles 28, Calif.

A meeting of the Steering Committee for the 1962 National Winter Convention on Military Electronics which will be held in Los Angeles, February 7–9, 1962. *Left to right:* Matthew E. Brady, Technical Program Chairman; Maynard D. McFarlane, Chairman of the Steering Committee; and Harry Delaney, Vice Chairman.

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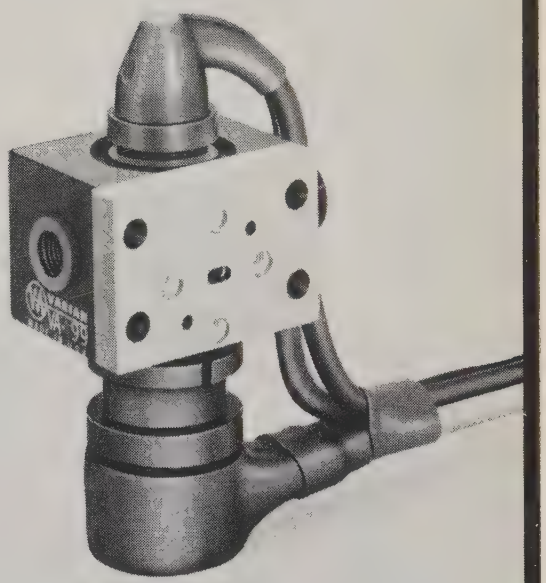

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Deadlines for 1962 JACC Announced

The 1962 Joint Automatic Control Conference will be held at New York University, University Heights, New York, on June 27–29, and will be jointly sponsored by the IRE-PGAC, AIEE, ISA, AIChE, and ASME. Innovations in paper review and documentation procedures are expected to result in a program of outstanding quality.

The control engineering faculty of the host university is headed by Dean John R. Ragazzini. The Conference General Chairman is Dr. Arthur S. Robinson, Director of Research, Kollsman Instrument Corp., and the Program Chairman is Anthony J. Hornfeck, Director of Research, Bailey Meter Co. The IRE representative on the Steering Committee is Prof. J. H. Mulligan, Electrical Engineering Dept., College of Engineering, New York University.

At the 1961 Joint Automatic Control Conference at Boulder, Colo., the 1962 Program Committee set the following paper deadlines for the 1962 Conference: abstracts by October 15, 1961, paper texts by November 15, 1961. Authors will be notified of paper disposition by February 15, 1962 and will have until April 15, 1962 to submit revised manuscripts. Prospective authors of significant papers covering control theory, applications or components are invited to submit abstracts and text through their sponsoring society. Early submission is encouraged.

Abstracts (4 copies) and complete papers (4 copies) from IRE authors should be submitted to the IRE-PGAC Program Committee Representative, Harold Levenstein, Servo Corp. of America, 111 New South Rd., Hicksville, Long Island, N. Y.

Operations Research and Management Sciences Groups to Meet

The first joint national meeting of the Operations Research Society of America and The Institute of Management Sciences will be held November 8–10, 1961, at the Jack Tar Hotel in San Francisco. The meeeting program will include sessions in three broad areas of Operations Research and Management Science. "Tools of OR/MS," under the chairmanship of Prof. George W. Brown, UCLA, will include mathematical programing, information processing, queuing models, and decision theory; Dr. Harold O. Davidson, Operations Research, Inc., is chairman for sessions on "The Applications of OR/MS," including management decision processes, problems in data and model validation, and case studies on military and nonmilitary applications. "The Challenge of OR/MS," with Prof. Gifford H. Symonds, Case Institute, as chairman will include use of OR/MS in the area of world peace, developments in organization theory, and general systems analysis. Both invited and contributed papers will be included in each session.

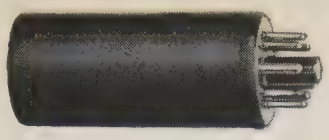
MISCELLANEOUS IRE PUBLICATIONS AVAILABLE

The following issues of miscellaneous publications are available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, New York, at the prices indicated below:

Meetings	Publications	Price per copy
Aeronautical and Navigational Electronics	*Proceedings of the 5th Annual East Coast ANE Conference,* held October 27–28, 1958 in Baltimore, Md.	$5.00*
Bio-Medical	*Digest of the 12th Annual Electrical Techniques in Medicine and Biology Conference,* held November 10–12, 1959 in Philadelphia, Pa.	4.00
	Digest of the 13th Annual Electrical Techniques in Medicine and Biology Conference, held October 31–November 2, 1960 in Washington, D. C.	5.00
	Digest of the 14th Annual Electrical Techniques in Medicine and Biology Conference and 4th International Conference on Medical Electronics, held July 16–21, 1961 in New York, N. Y.	5.00
	Bibliography on Medical Electronics June 1958 issue	2.50
	Bibliography on Medical Electronics, June 1959 issue (Supplement I)	2.50
	Bibliography on Medical Electronics, July 1960 issue (Supplement II)	2.50
Communications	*Record of the 5th Global Communications Symposium, (GlobeCom V),* held May 22–24, 1961 in Chicago, Ill.	4.00
Computers	*Proceedings of the Joint AIEE IRE ACM Eastern Electronic Computer Conference,* held December 10–12, 1951 in Philadelphia, Pa.	3.50
	Proceedings of the Joint AIEE IRE ACM Eastern Electronic Computer Conference, held December 8–10, 1954 in Philadelphia, Pa.	3.00
	Proceedings of the Joint AIEE IRE ACM Eastern Electronic Computer Conference, held November 7–9, 1955 in Boston, Mass.	3.00
	Proceedings of the Joint AIEE IRE ACM Eastern Electronic Computer Conference, held December 13–15, 1960 in New York, N. Y.	3.00
	Proceedings of the Joint AIEE IRE ACM Western Electronic Computer Conference, held February 7–9, 1956 in San Francisco, Calif.	3.00
	Proceedings of the Joint AIEE IRE ACM Western Electronic Computer Conference, held May 9–11, 1961 in Los Angeles, Calif.	4.00
Magnetic Amplifiers	*Proceedings of the Magnetic Amplifiers Conference,* held April 5–6, 1956 in Syracuse, N. Y.	4.00
Military Electronics	*Proceedings of the 1st National Military Electronics Convention,* held June 17–19, 1957 in Washington, D. C.	5.00†
	Proceedings of the 2nd National Military Electronics Convention, held June 16–18, 1958 in Washington, D. C.	5.00†
	Proceedings of the 3rd National Military Electronics Convention, held June 29–July 1, 1959 in Washington, D. C.	4.00
	Proceedings of the 4th National Military Electronics Convention, held June 27–29, 1960 in Washington, D. C.	5.00
	Proceedings of the 5th National Military Electronics Convention, held June 26–28, 1961 in Washington, D. C.	5.00

Continued on page 20A

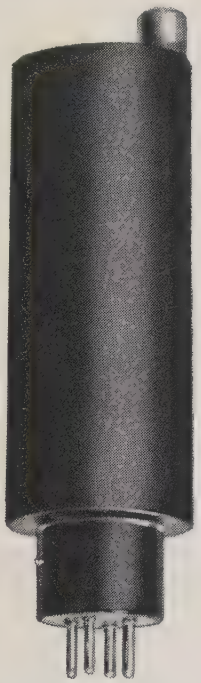
* IRE Member rate $3.50.
† IRE Member rate $3.00.



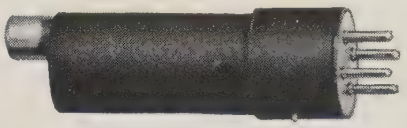
S-5018—replaces 5AU4, 5AW4, 5AZ4, 5T4, 5U4, 5V4 5W4, 5Y3, 5Z4, 5931, 6087, 6106

S-5207
replaces 6X4

S-5130—replaces 866, 866A, 3B28

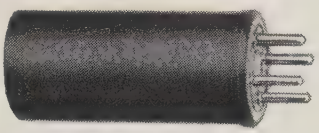


S-5367—replaces 8020

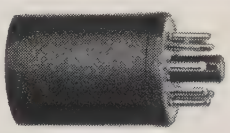


S-5343—replaces 816, 836, or 3B28 and 866 at reduced voltage

S-5033—replaces 6AU4, 6AX4, 6BL4, 6W4, 12AX4, 17AX4, 25W4, 6U4
S-5019—replaces 5R4



S-5011A—replaces 80, 82, 83, 83V, 5Z3
S-5017—replaces 0Z4, 5X4, 5Y4, 6AX5, 6X5



S-5251—replaces 5U4, 5AU4, 5AW4, 5AZ4, 5T4, 5V4, 5W4, 5Y3, 5Z4

Illustrations approximately ½ size

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S-5344
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Conference Location Sought by JECMB

The Joint Executive Committee on Medicine and Biology at a recent meeting approved the following resolution:

"The JECMB committee is soliciting invitations for future conference locations. An offer has been received from the Baltimore, Md. PGBME Section of the IRE to hold the 1963 conference in that city. Other bids for the 1963 conference location will be considered if received before November 1, 1961."

Communications should be addressed to Dr. Robert L. Bowman, Chairman, JECMB, National Heart Institute, National Institute of Health, Bethesda 14, Md.

Obituaries

William A. MacDonald (A'19–M'26–SM'43–F'53) died on August 11, 1961 at his home in Locust Valley, Long Island, N. Y. He was born in New York City on December 27, 1895.

W. A. MacDonald

He began his career in radio and electronics as a radio operator for the Marconi Wireless and Telegraph Company; from 1916 to 1917, he worked for the Western Electric Company in the Physical Laboratory. He served in the Army Signal Corps during World War I and worked in Paris with Major E. H. Armstrong on the first superheterodyne radio receiver. After his release from active duty, he worked as an engineer for the Signal Corps and established what became the Wright Field Radio Laboratory at Dayton, Ohio. He then worked in the Consulting Engineering Department of the Radio Corporation of America from 1920–1924.

In 1924, he joined the Hazeltine Corporation as Chief Engineer where he was responsible for the staff that developed inventions for the radio industry and later for monochrome and color television. He was a director and vice president of the corporation at the beginning of World War II when he steered the company into military electronics. He was elected President of Hazeltine in 1952 and Chairman of the Board in 1957. Earlier this year, he was elected a director of the Western Union Telegraph Company.

A charter member of the Electronic Equipment Advisory Committee of the Munitions Board and a member of the Joint Electronics Industry Committee, he also served on the Department of Defense Advisory Group on Reliability of Electronics Equipment Task Group 6. For many years, he was a director of the Electronic Industries Association.

He was a member of the Armed Forces Communications and Electronics Association, the American Society of Naval Engineers, American Ordnance Association and

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Radio Frequency Interference	*Digest of the 2nd National RFI Symposium,* held June 13–14, 1960 in Washington, D. C.	$2.00
	Digest of the 3rd National RFI Symposium, held June 12–13, 1961 in Washington, D. C.	4.00
Reliability and Quality Control in Electronics	*Proceedings of the 1st National RQC Symposium,* held November 12–13, 1954 in New York, N. Y.	5.00
	Proceedings of the 2nd National RQC Symposium, held January 9–10, 1956 in Washington, D. C.	5.00
	Proceedings of the 3rd National RQC Symposium, held January 14–16, 1957 in Washington, D. C.	5.00
	Proceedings of the 4th National RQC Symposium, held January 6–8, 1958 in Washington, D. C.	5.00
	Proceedings of the 5th National RQC Symposium, held January 12–14, 1959 in Philadelphia, Pa.	5.00
	Proceedings of the 6th National RQC Symposium, held January 11–13, 1960 in Washington, D. C.	5.00
	Proceedings of the 7th National RQC Symposium, held January 9–11, 1961 in Philadelphia, Pa.	5.00
Telemetering	*Proceedings of the 1959 National Space Electronics and Telemetry Symposium,* held September 28–30, 1959 in San Francisco, Calif.	5.00
	Proceedings of the 1960 National Telemetering Conference, held May 23–25, 1960 in Santa Monica, Calif.	6.50‡
	Proceedings of the 1961 National Telemetering Conference, held May 22–24, 1961 in Chicago, Ill.	3.50

‡ IRE Member rate $4.50.

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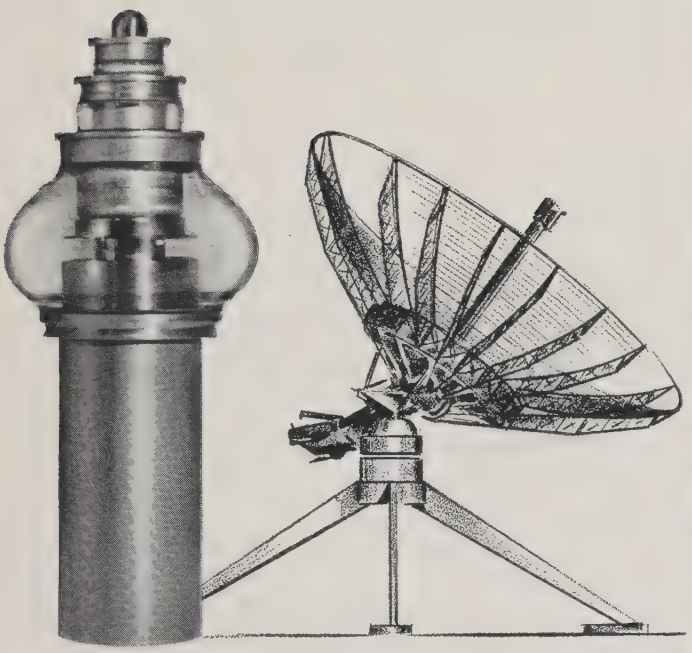

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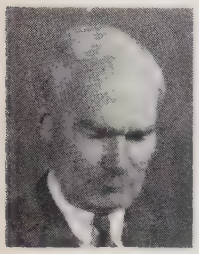
National Security Industrial Association. He was also a licenced professional engineer and a Fellow of the Radio Club of America, a founding member of the Armstrong Research Foundation and a member of the AIEE.

Mr. MacDonald was granted 110 United States and foreign patents and was awarded a Presidential Citation for his contributions to electronics during World War II.

With the recent passing of Lee de Forest (A'12–M'14–F'18), the electronics profession lost a renowned pioneer and inventor, and the IRE a distinguished charter member.



L. de Forest

His invention of the three-element audion tube in 1906 provided a cornerstone for the foundation of the communications field, earning him the title, "father of radio." In the course of his long and fruitful career, during which he was awarded more than 300 patents, he took an active part in the development of radar, television, sound motion pictures, long-distance telephones, cosmic ray measurements, electric phonographs and diathermy machines.

President of the Society of Wireless Telegraph Engineers in 1909–1910, Dr. de Forest was one of the original group of 46 members who participated in the merger of the STWE and the Wireless Institute to form the IRE in 1912. He served as President of the IRE in 1930 and Director from 1930 to 1932.

Lee de Forest was born in Council Bluffs, Iowa, on August 26, 1873. In 1893 he entered the Sheffield Scientific School at Yale University, where he enrolled in the first course in electricity to be offered by an American college. He was graduated with the B.S. degree in 1896 and took his Ph.D., also at Yale, three years later.

His first job was with the Western Electric Company in Chicago where, with the collaboration of Edward Smythe, he invented a radio receiver using an electrolytic detector that made possible the use of headphones for wireless messages. This was his first patent.

He soon helped to form the American de Forest Wireless Telegraph Company in New York City. The firm went bankrupt in 1906, and Dr. de Forest then organized the de Forest Radio Telegraph Company in 1907. In 1909, the firm merged with North American Wireless Corp. and de Forest went to work for the Federal Telegraph Company in Palo Alto, Calif.

Before moving to California, he conducted the first radio broadcast of live music by erecting a radio-telephone transmitter at the Metropolitan Opera House and placing two microphones on the stage. The occasion was the opening night of Enrico Caruso in "Cavalleria Rusticana" in 1909. The broadcast was heard by only a handful of de Forest's friends at his home and a few amateurs and ship wireless operators.

Making California his permanent home, he spent the remainder of his career there, producing many more inventions. His last patent, for an improvement in magnetic tape, came in 1958.

During his career he experienced several financial reverses and spent considerable time and money in long-drawn-out court battles over his inventions. His most controversial lawsuits included litigations with the Marconi Company, Irving Langmuir, Major E. H. Armstrong, and German inventors of talking motion pictures.

He received many honors during his lifetime including the IRE Medal of Honor in 1922, the Edison Medal of the American Institute of Electrical Engineers, the Elliott Cresson Medal of the Franklin Institute, the John Scott Medal, and an award from the National Association of Broadcasters. He also received the Cross of the Legion of Honor from the French Government and honorary degrees of Doctor of Science from Syracuse and Yale Universities.

National Symposium on Engineering Writing and Speech

Kellogg Center for Continuing Education, Michigan State University, East Lansing, October 16–17, 1961

The National Symposium on Engineering Writing and Speech, which will be held at the Kellogg Center for Continuing Education, East Lansing, Mich., on October 16–17, 1961, has as its theme, "Communication: The Modern Engineer's Function." The program has been announced as follows:

Monday Morning, October 16

Keynote Address,* *Dr. D. E. Gray, Natl. Science Liaison Officer, Natl. Science Foundation, Washington, D. C.*

"The Purpose and Function of PGEWS,"* *E. McElwee, Harrison Labs., RCA, Harrison, N. J.*

"Hellbent to Communicate,"* *M. Brotherton, Bell Telephone Labs., New York, N. Y.*

"Sign Language in Engineering—Asset or Liability," *L. D. Kovach, Dept. of Mathematics and Physics, Pepperdine College, Los Angeles, Calif.*

"Mathematics in Technical Communication,"* *H. E. Koenig, Dept. of Elec. Engrg., Michigan State University, East Lansing, Mich.*

Monday Luncheon

Address,* *J. D. Ryder, Dean of Engineering, Michigan State University, East Lansing, Mich.*

Monday Afternoon

"Making Visual Communications Communicate,"* *D. M. Davis, Visual Communications Div., Office of Administrative Services, Wright-Patterson Air Force Base, Ohio.*

Workshop on Visual Communications, *Tecnifax Corp., Holyoke, Mass.*

Monday Dinner

Address,* *F. Leamer, Personnel Director, Bell Telephone Labs., Murray Hill, N. J.*

Tuesday Morning, October 17

"The Self-Organizing System and Its Ability to Communicate,"* *B. F. Chaydleur, Computer Div., Philco Corp.*

"Information Identification—The Challenge of Technical Communication," *W. A. Ternent, Information Services, Defense Systems Dept., General Electric Co., Syracuse, N. Y.*

* Invited paper.

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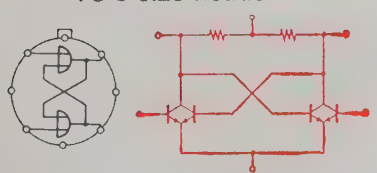

"G" GATE
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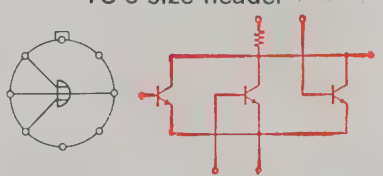

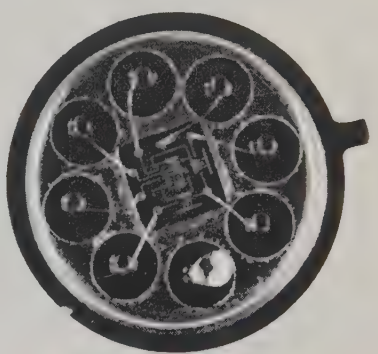

"S" HALF-SHIFT REGISTER
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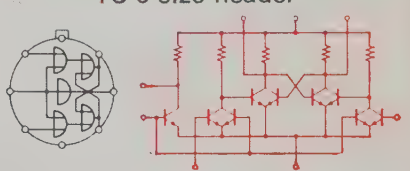

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"Communications for the Engineering Manager," *J. B. Bennett, Research Directorate, Systems Development Corp., Santa Monica, Calif.*

"1100 Prospective Authors," *M. Ellis, Customer Relations Div., Aerojet-general Corp., Sacramento, Calif.*

"Communicating Ideas: The Modern Engineer's Function," *W. J. Van De Mark, Bell Telephone Labs., Whippany, N. J.*

Tuesday Luncheon

Address,* *W. L. Dunn, Vice President and Director of Engineering and Research, Admiral Corp., Chicago, Ill.*

Tuesday Afternoon

"The New Look in Military Marketing Literature," *R. E. Hohmann, IBM Command Control Center, Federal Systems Div., Kingston, N. Y.*

"The Role of Language in Military Systems Proposals," *N. Maroulis, Light Military Electronics Dept., General Electric Co., Utica, N. Y.*

"Proposals with a Future," *J. Puopolo, Light Military Electronics Dept., General Electric Co., Utica, N. Y.*

Workshop on Writing,* *conducted by R. C. Gerfen, Dept. of Business Writing, Northwestern University, Evanston, Ill.*

Symposium on Instrumentation Facilities for Biomedical Research

SHERATON-FONTENELLE HOTEL, OMAHA, NEB., OCTOBER 26–27, 1961

This Conference on "The Role of Biomedical Engineering in Universities and Hospitals" will study the present and future role of biomedical engineering as a scientific resource. It will bring together representatives from universities, research institutions, hospitals and industry. The principal feature of the conference will be four study groups meeting simultaneously to discuss major problems in the development of this field. All registrants will be assigned to the study group of their choice and will be encouraged to participate in the discussions. The work of these groups will be summarized and reported at the final session. It is hoped that the conference proceedings, which will include the reports of the study groups, will serve as a guide to further development of biomedical engineering in universities and hospitals.

The fee of $13.00 includes all sessions and lunch and dinner Thursday, October 26, 1961. Registration is from 8:00–10:00 P.M. October 25 and 8:00 A.M. October 26 at the Sheraton-Fontenelle Hotel. The provisional program follows:

Thursday Morning, October 26

Session I—The Role of Engineering in Biological and Medical Research

Welcome: Session Chairman: *F. L. Dunn, University of Nebraska.*

"Biomedical Engineering in Basic Research," *W. P. Peak, Mass. Inst. Tech.*

"Biomedical Engineering in Clinical Medicine," *D. Holaday, University of Chicago.*

"Industry's Contribution to Biomedical Education and Research," *A. Grass, Grass Instrument Co.*

"The Relationship between Bionics and Biomedical Engineering," *J. Lyman, University of California, Los Angeles.*

"The Impact of Engineering Theory on Biology," *R. L. Schoenfeld, Rockefeller Institute and Polytechnic Institute of Brooklyn.*

Thursday Afternoon

Session II—The Study Groups

The four study groups will convene simultaneously. A summary of the conclusions of each group will be presented by the chairman during the last session of the conference.

Group A—Development of the Biomedical Engineer

Discussions centered around:
1) Scope of scientific training
2) Development of professional maturity in research
3) Problems of professional growth

Discussion leaders: *V. W. Bolie, Chairman, Iowa State University; E. Gordy, Roswell Park Memorial Inst.; J. Lyman, University of California, Los Angeles; L. H. Montgomery, Vanderbilt University; S. A. Talbot, Johns Hopkins University.*

Group B—Research Organization of Biomedical Engineering

1) The research project
2) Departmental programs
3) Institutional facilities

Discussion leaders: *R. L. Schoenfeld, Chairman, Rockefeller Institute; J. Artesani, Wake Forest College; H. Fein, University of Utah; R. E. Jones, Mayo Clinic; G. C. Riggle, National Institutes of Health.*

Group C—Development of Research Instrumentation

1) Problems in specifications and methodology
2) Evolution from design

Discussion leaders: *S. R. Gilford, Chairman, Gilford Instruments; D. Baker, University of Washington; A. Grass, Grass Instrument Co.; G. I. Johnston, University of Oregon; W. Mueller, State University of New York.*

Group D—Administrative Organization

1) Relation to research function
2) Problems of financing
3) Staff structure

Discussion leaders: *G. N. Webb, Chairman, Johns Hopkins Hospital; R. L. Bowman, National Institutes of Health; F. L. Dunn, University of Nebraska; C. Schaeffer, Albany Medical College; R. Stacy, Ohio State University.*

Thursday Evening

Informal Conference Session

Social hour, informal introduction of committee members and dinner.

Friday Morning, October 27

Session III—Support Programs for Biomedical Engineering

Further discussion group meetings, reconsideration of study group reports.

Session Chairman: *G. C. Riggle, National Institutes of Health.*

"Current Status of Educational Programs," *F. L. Thurstone, North Carolina State College.*

"Engineering-SciencesPrograms," (*Speaker to be announced*) *National Science Foundation.*

"Industrial Aid to Biomedical Engineering," *L. E. Slater, Foundation for Instrumentation Education and Research.*

"Research and Training Grant Program Relating to Biomedical Engineering," *W. L. Koltun, National Institutes of Health.*

(Title to be announced) *M. McLennan, N.A.S.A. and U.S.A.F. Air Medical Group.*

Friday Afternoon

Session IV—Reports from the Study Groups

Session Chairman: *D. LeCroisette, Drexel Institute of Technology.*

Study Group Chairmen: Group A: *V. W. Bolie*; Group B: *R. L. Schoenfeld*; Group C: *S. R. Gilford*; Group D: *G. N. Webb.*

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1961 Electron Devices Meeting

SHERATON PARK HOTEL, WASHINGTON, D.C., OCTOBER 26–28, 1961

Thursday, October 26

Session 1(A)—High Power Microwave Tube

Chairman: *M. Chodorow, Stanford University, Stanford, Calif.*

Organizer: *L. S. Nergaard, RCA Labs., Princeton, N. J.*

"A Broadband Megawatt Traveling Wave Tube Utilizing a Ring-Loaded Meander Line Interaction Circuit," *C. Burklund, Sperry Gyroscope Co., Great Neck, N. Y.*

"A 5 Megawatt-300 Kilowatt Klystron Amplifier at L-Band," *R. E. Holaday and G. W. Peterson, Litton Industries, San Carlos, Calif.*

"A 1.5 Megawatt, High Efficiency, X-Band CEM Coaxial Magnetron," *A. Cook, J. Drexler, F. Feulner, S-F-D Labs., Inc., Union, N. J.*

"A 500-KW Periodically Focused S-Band Traveling-Wave Amplifier," *B. A. Highstrete and A. L. Rousseau, Hughes Aircraft Co., Los Angeles, Calif.*

"Experimental Broadband Klystrons Using High Perveance Hollow Beams," *R. E. Holaday and L. Fox, Litton Industries, San Carlos, Calif.*

"Experimental Results on a High Power CW Klystron with an Extended Interaction Catcher," *D. Preist and W. J. Leidigh, Eitel-McCullough, Inc., San Bruno, Calif.*

Session 1(B)—Transistors

Chairman: *R. L. Pritchard, Texas Instruments, Inc., Dallas, Tex.*

Organizer: *I. A. Lesk, Motorola, Inc., Phoenix, Ariz.*

"Germanium Transistors for Operation above 1 KMC," *J. T. Nelson and A. G. Foyt, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"KMC Silicon Planar Transistors," *D. F. Allison, O. Baker, and G. E. Moore, Fairchild Semiconductor, Palo Alto, Calif.*

"Design of a High Frequency Transistor with Star Geometry," *J. C. Haenichen, T. R. Huffman, J. E. Muschinske, and P. T. Robinson, Motorola, Inc., Phoenix, Ariz.*

"A Transistor Utilizing an Epitaxially Grown Base and Collector Region," *J. K. Clifton and H. M. Robertson, Motorola, Inc., Phoenix, Ariz.*

"Planar Germanium Diodes and Transistors," *K. Simonyan and P. Constantakes, General Instrument Corp., Hicksville, L. I., N. Y.*

"An NPN Germanium Double-Diffused Mesa Transistor," *R. A. Jacobs, R. H. Abshire, R. W. Wilson and H. Anetsman, Motorola, Inc., Phoenix, Ariz.*

"A Germanium Ultra High Speed Switching Transistor," *D. Granberry, D. D. Martin, E. Orris and C. M. Chang, Texas Instruments, Inc., Dallas, Tex.*

"Injected Carrier Concentrations in Diffused Silicon Transistors by Observations of Interband Carrier Recombination," *N. J. Chaplin, Bell Telephone Labs., Inc., Allentown, Pa.*

Session 1(C)—Integrated Electronic Devices

Chairman: *H. Moss, Texas Instruments, Inc., Dallas, Tex.*

Organizer: *J. Nall, Fairchild Semiconductor Corp., Palo Alto, Calif.*

"The Unibi Functional Block—A Monolithic Amplifier with Unipolar-Bipolar Structure," *K. Yu and L. Pollock, Westinghouse Electric Corp., Pittsburgh, Pa.*

"Solid State Micrologic Elements," *N. Gault, L. Kattner, and J. Nall, Fairchild Semiconductor Corp., Palo Alto, Calif.*

"Tunable High Q Solid State Inductive Elements," *C. Fa, H. Dill and W. Waters, Hughes Aircraft Co., Newport Beach, Calif.*

"Maximum Packing Density and Minimum Size of Semiconductor Devices," *J. T. Wallmark and S. M. Marcus, RCA, Princeton, N. J.*

"Semiconductor Delay Line Tuner," *J. S. Winslow, Electro-Optical Systems, Inc., Pasadena, Calif.*

Session 1(D)—Energy Sources

Chairman: *W. G. Evans, Westinghouse Electric Corp., Pittsburgh, Pa.*

Organizer: *T. M. Corry, Westinghouse Electric Corp., Pittsburgh, Pa.*

"High-Efficiency Gallium-Arsenide Solar Cells," *D. P. Bortfeld, A. R. Gobat, M. F. Lamorte, and G. W. McIver, RCA, Somerville, N. J.*

"Solar Power Conversion Using Photoelectric Emission," *A. S. Jensen and I. Limansky, Westinghouse Electric Corp., Baltimore, Md.*

"Radiation Resistant Solar Cells for Communication Satellites," *K. D. Smith, W. G. Ansley, D. F. Ciccolella, and M. P. Lepselter, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Thermoelectric Generators—Design, Performance and Application," *T. M. Corry and G. Spira, Westinghouse Electric Corp., Pittsburgh, Pa.*

"Design and Development of a High Power Density Portable Thermoelectric Generator," *W. C. Moreland and J. C. Kastovich, Westinghouse Atomic Equipment, New Products Lab., Cheswick, Pa.*

"The Performance Limits and Design Techniques of Multistage Thermoelectric Devices," *I. I. Sochard, Diamond Ordnance Fuze Labs., Washington, D. C.*

Friday, October 27

Session 2—Invited Papers

Chairman: *W. Adcock, Texas Instruments, Inc., Dallas, Tex.*

Organizer: *I. M. Ross, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Masers and Optical Masers," *R. Kompfner, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Whither the Tunnel Diode?" *E. O. Johnson, RCA, Somerville, N. J.*

"Microminiaturization—Problems it can Solve and Problems it can Create," *R. H. Baker, M.I.T., Lincoln Lab., Lexington, Mass.*

Session 3(A)—Storage and Display Tubes

Chairman: *W. G. Dow, University of Michigan, Ann Arbor, Mich.*

Organizer: *R. C. Hergenrother, Raytheon Mfg. Co., Waltham, Mass.*

"A New Device for Radar Integration," *R. C. Wilcox, RCA, Van Nuys, Calif.*

"A Display for 100 Million Dots on a Five-Inch Face," *K. Schlesinger, General Electric Co., Syracuse, N. Y.*

"A Scientific Explanation of the "Self-Erasing" or "Decay to Black" Phenomenon in Direct-Viewing Storage Tubes," *V. LeGendre and R. Madden, Allen B. Du Mont Labs., Clifton, N. J.*

"Two-Color Display Storage Tube," *C. D. Beintema, N. J. Koda and L. S. Yaggy, Hughes Aircraft Co., Oceanside, Calif.*

"Trends in Gas-Filled Trigger Tube Development," *S. M. Frouws, N. V. Phillips' Gloeilampenfabrieken, Eindhoven, Holland.*

"A Cathode-Ray Switch Tube With 10 Channels of Electron Multiplication," *R. Kalibjian, Lawrence Radiation Lab., University of California, Livermore, Calif.*

"A Grid Controlled Cold Cathode Visual Indicator," *P. W. Stutsman and N. E. Linson, Raytheon Co., Newton, Mass.*

Session 3(B)—Diodes

Chairman: *W. C. Dunlap, Raytheon Co., Waltham, Mass.*

Organizer: *J. M. Early, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Broad Aspects of the Interaction in Parametric Devices," *H. J. Evans and H. Hsu, General Electric Co., Syracuse, N. Y.*

"High-Cutoff-Frequency Gallium-Arsenide Diffused-Junction Varactor Diodes," *L. H. Gibbons, M. F. Lamorte and A. Widmer, RCA, Somerville, N. J.*

"Tunnel Diodes for Low Noise Amplification," *L. Armstrong, Micro State Electronics Corp., Murray Hill, N. J.*

"Fabrication of a Rugged, Low-Peak Current High-Speed Tunnel Diode," *S. S. Im and J. S. Logan, Internatl. Business Machines Corp., Poughkeepsie, N. Y.*

"Hermetically Sealed Silicon Chip Diodes and Transistors," *J. L. Langdon, W. E. Mutter, R. P. Pecoraro, K. K. Schuegraf, Internatl. Business Machines Corp., Poughkeepsie, N. Y.*

"Hot-Glass Sealed Silicon Diodes," *V. Decker, J. H. Forster, B. T. Howard, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"PN Junctions and Solar Cells by Ion Implantation," *W. J. King and S. J. Solomon, Goodrich High Voltage Astronautics, Inc., Burlington, Mass.*

"The Significance of Crystallographic Polarity in the Fabrication of Alloyed Junctions in InSb," *M. T. Minamoto, Airborne Instruments Lab., Deer Park, N. Y.*

Session 3(C)—New Tubes and Theory

Chairman: *R. Hutter, Sylvania Electric Products, Inc., Mountain View, Calif.*

Organizer: *D. A. Chisholm, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Microwave High Power Nanosecond Switch Using Multipactor Discharge," *C. Milazzo, General Electric Co., Palo Alto, Calif.*

"The Dematron—A New Crossed-Field Amplifier," *G. E. Pokorny, A. E. Kushnick and J. F. Hull, Litton Electron Tube Corp., San Carlos, Calif.*

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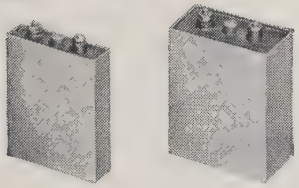

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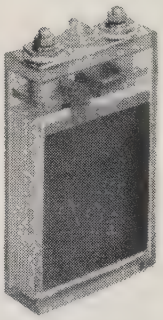

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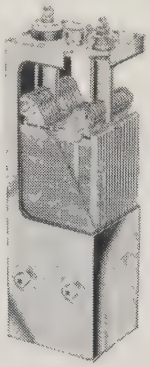

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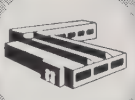


Switchgear Operation
Emergency Lighting
Engine Starting
Flashlights and Lanterns
Cordless Power Tools

MARINE

Engine Starting
Radio and Alarms
Life Boats

"Broadband, High Power, Coaxial Arc Discharge Switch," *S. J. Tetenbaum, D. L. Campbell and R. R. Moats, Sylvania Electric Products, Inc., Mountain View, Calif.*

"The Mercury-Arc Oscillator," *W. D. Jackson and T. K. Chiang, Mass. Inst. Tech., Cambridge, Mass.*

"The Effect of Helix Loss on Noise Figure in Traveling-Wave Tubes," *B. P. Israelsen, Watkins-Johnson Co., Palo Alto, Calif.*

"The Calculation of Accurate Triode Characteristics Using a Modern High-Speed Computer," *O. H. Schade, Sr., RCA, Harrison, N. J.*

Saturday, October 28

Session 4(A)—CW Microwave Tubes

Chairman: *G. Wade, Raytheon Co., Spencer Lab., Burlington, Mass.*

Organizer: *S. Bloom, RCA Labs., Princeton, N. J.*

"An X-Band Traveling-Wave Tube for Space Applications," *J. Herman, Sperry Gyroscope Co., Great Neck, N. Y.*

"A 4 KMC Satellite TWT for the AT&T Communication Satellite," *M. G. Bodmer, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Electron Tubes at Millimeter and Submillimeter Wavelengths," *Yéou-Ta and G. Convert, Compagnie Générale de Telegraphie Sans Fil, Paris, France.*

"A Non-Degenerate Electron Beam Parametric Amplifier," *G. W. Hrbek, Zenith Radio Corp., Chicago, Ill.*

"A 12-Watt Traveling-Wave Tube Amplifier for Space Communications," *J. N. Nelson and L. A. Roberts, Watkins-Johnson Co., Palo Alto, Calif.*

"Novel K-Band Reflex Klystrons with High Degree of Frequency-Temperature Stability," *R. G. Rockwell, Varian Associates, Palo Alto, Calif.*

Session 4(B)—Special Solid State Devices

Chairman: *W. M. Webster, RCA Labs., Princeton, N. J.*

Organizer: *J. Hilibrand, RCA Labs., Princeton, N. J.*

"Design and Application of Silicon Unipolar Transistors Made by Diffusion Techniques," *W. A. Bösenberg, J. A. Olmstead, and K. Wybrands, RCA, Somerville, N. J.*

"The Cryosistor—A Field-Effect Controlled Impact Ionization Switch," *I. Melngailis, Carnegie Inst. Tech., Pittsburgh, Pa.*

"Conduction Through Thin Metal, Metal Oxide Structures," *L. E. Godycki, D. P. Foote and I. Weiman, Electro-Optical Systems, Inc., Pasadena, Calif.*

"A Multi-Position Magnetic Core Driver Using Electron-Beam Switched p-n Junctions," *A. V. Brown, I.B.M., Yorktown Heights, N. Y.*

"Germanium PME Fast Response Light Detector," *A. Boatright, H. Mette and E. R. Ahlstrom, U. S. Army Signal Research and Development Lab., Fort Monmouth, N. J.*

"Electroluminescent Device for Temporary Information Storage," *J. W. Culp, General Telephone and Electronics Labs., Inc., Bayside, N. Y.*

Session 4(C)—Selected Tube Topics

Chairman: *J. A. Hornbeck, Bell Telephone Labs., Inc., Murray Hill, N. J.*

Organizer: *F. E. Vaccaro, RCA, Princeton, N. J.*

"Effects of Velocity Tapering and D-C Gradients in Traveling-Wave Tubes," *J. G. Meeker and J. E. Rowe, University of Michigan, Ann Arbor, Mich.*

"Cathodes with Smooth Oxide Coatings," *H. J. Schuetze and W. E. Dahlke, Telefunken, Ulm (Donau), Germany.*

"Power Handling Ability of Helices Mounted in Beryllium Oxide," *H. K. Detweiler and J. E. Rowe, University of Michigan, Ann Arbor, Mich.*

"An Electron-Beam-Probe for Potential Measurements in Plasmas," *R. F. Kemp, J. C. Slattery and J. M. Sellen, Jr., Space Technology Labs., Inc., Canoga Park, Calif.*

"The Dark Heater Improves Tube Reliability," *E. M. Troy, RCA, Harrison, N. J.*

"Investigation and Evaluation of Two New Non-Planar Ladder Derived Delay Lines," *H. J. Krahn, Raytheon Co., Burlington, Mass.*

"Ray Tracing on the Electron Optica Bench," *M. P. Siedband and R. A. Yeates Westinghouse Electric Corp., Baltimore, Md.*

1961 Radio Fall Meeting

HOTEL SYRACUSE, SYRACUSE, N. Y., OCTOBER 30–NOVEMBER 1, 1961

The tentative program for the 1961 Radio Fall Meeting to be held October 30 to November 1, 1961 at the Hotel Syracuse, Syracuse, N. Y. has been announced. It is as follows:

Monday Morning, October 30
Registration

EIA Engineering Department Session

Chairman: *J. R. Whitney*

"Recent Changes in the JEDEC Type Designation System," *G. F. Hohn, EIA Standards Lab.*

"The Significance of the EIA Music Power Output Rating System on Electron Tube and Transmission Amplifiers," *J. A. Stark, General Electric Co.*

Monday Afternoon

Reliability and Quality Control Session

(Arranged by the IRE Professional Group on Reliability and Quality Control.)

Chairman: *D. Kerins*

"A Comparison of Commercial and Military Reliability Problems," *A. P. Mundel, Sonotone, Inc.*

"Application of Reliability Engineering and Statistics Techniques in the Development of Commercial Product Warrantees," *G. R. Herd, Booz, Allena and Hamilton.*

"Potential Application of Reliability Techniques in Communication Product Lines," *R. M. Jacobs, Sylvania Electric Products, Inc.*

Tuesday Morning, October 31

Electron Tube Session

(Arranged by the IRE Professional Group on Electron Devices.)

Chairman: *A. Szilasi*

"Design Considerations of Novar Tube Types," *M. Bondy, Radio Corp. of America.*

"Progress in Video Amplifier Tubes for TV Receivers," *A. J. Robinson and R. F. Bergdahl, Sylvania Electric Products, Inc.*

" Low Noise VHF Collimated Beam Tetrode," *N. Buchler and R. George, Tung-Sol Electric, Inc.*

"An Ultra-Linear Transfer Characteristic," *A. vanderJagt, Westinghouse Electric Corp.*

"The Improved Sweep System for the 90° Family of Cathode Ray Tubes," *H. C. Klender, National Video Corp. and K. McGlashan, Advance Ross Electronics Corp.*

Tuesday Afternoon

Semiconductor Devices Session

(Arranged by the IRE Professional Group on Electron Devices.)

Chairman: *E. Johnson*

"Variation of High Frequency Power Gain and Noise Figures vs Temperature," *C. R. Gray and T. C. Sowers, Lansdale Div., Philco Corp.*

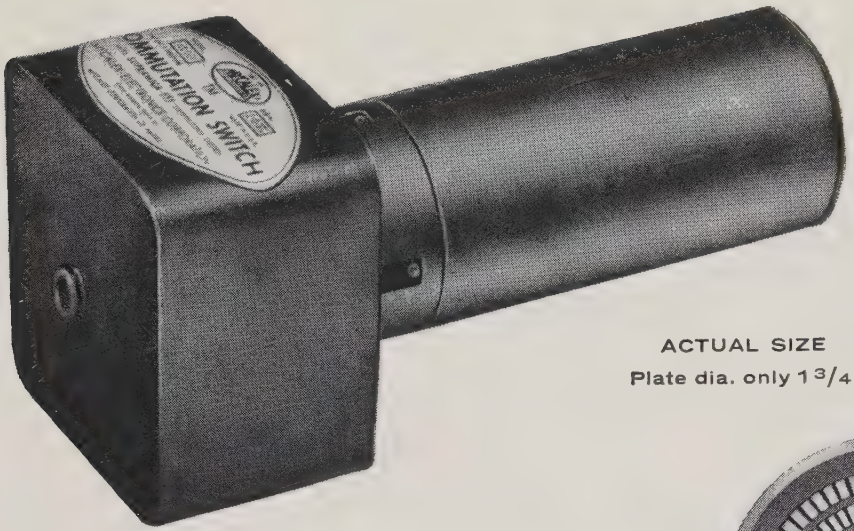
"Transistorized Horizontal Deflection Circuits," *R. Hartz, Bendix Corp.*

"An All Transistor Stereo Hi-Fi System," *D. Taylor, Motorola, Inc.*

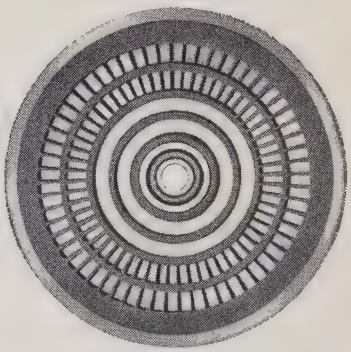
"A New High Performance Transistor Complement for TV Receivers," *B. M. Soltoff and C. D. Simmons, Lansdale Div., Philco Corp.*

"Consideration of FM Receiver Design for Mobile Applications," *H. Thanos, RCA, Sommerville, N. J.*

NEW MINIATURIZED LONG-LIFE COMMUTATION SWITCH



ACTUAL SIZE
Plate dia. only 1 3/4"



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Tuesday Evening
Radio Fall Meeting Dinner

Toastmaster: *F. R. Lack, Director, EIA Engrg. Dept.*
Speaker to be announced.

Wednesday Morning, November 1
Radio and Television Session I

(Arranged by the IRE Professional Group on Broadcast and Television Receivers.)
Chairman: *R. R. Thalner*
"A Survey of the Performance Characteristics of Television Receivers and Antennae for the UHF-TV Project," *E. W. Chapin, Federal Communications Commission.*
"An Airborne UHF Television Transmitter for Educational TV," *F. Mullin, Westinghouse Electric Corp.*
"Complementary Symmetry Sliding Class-A Audio System," *Z. Wiencek and J. Chua, Warwick Manufacturing Corp.*
"Specification of Transistor Parameters for Broadcast Receivers," *M. Slavin and H. J. Laurent, Bendix Corp.*

Wednesday Afternoon
Radio and Television Session II

(Arranged by the IRE Professional Group on Broadcast and Television Receivers.)
Chairman: *D. E. Harnett*
"Thoughts on the Subjective Aspects of Stereo Reproduction," *C. J. Hirsch, Radio Corp. of America.*
"Design Considerations in the Development of a High Quality AM-FM Console for Stereo Multiplex," *Zenith Radio Corp.*
"A Stabilized Ultrasonic Electrostatic Transducer," *W. L. Hughes, General Electric Co.*
"The Effects of Reduced Heater Power Standby Operation on the Reliability of Television and Radio Receivers," *C. B. Heffron, Westinghouse Electric Corp.*

Northeast Electronics Research and Engineering Meeting

COMMONWEALTH ARMORY AND SOMERSET HOTEL, BOSTON, MASS., NOVEMBER 14–16, 1961

The NEREM 1961 technical program will include over 90 papers written by more than 100 scientists and engineers. The keynote address will be given on Tuesday, November 14, by John L. Burns, President of RCA, and a major address at the annual banquet on Wednesday, November 15, by Charles H. Townes, Provost of MIT. At the keynote address ceremony, K. C. Black, General Chairman of NEREM, will officiate, and John A. Volpe, Governor of the Commonwealth of Massachusetts, will present a welcoming address.

Tuesday Afternoon, November 14

Session I—Stereo Broadcasting

Chairman: *D. Von Recklinghausen, H. H. Scott, Inc., Maynard, Mass.*
"FM Broadcasting Standards as Evolved Through History," *H. L. Kassens, FCC, Washington, D. C.*
"Conversion of Existing FM Transmitters for Stereo Transmission," *F. J. Bias, General Electric Co., Syracuse, N. Y.*
"Monitoring of FM Stereophonic Signals," *A. J. DeVries, Zenith Radio Corp., Chicago, Ill.*
"Design Aspects of FM Stereo-Tuners and Adapters," *D. Von Recklinghausen and H. H. Scott, H. H. Scott, Inc., Maynard, Mass.*

Session II—Region I Student Prize Papers

Chairman: *R. E. Gildersleeve, Syracuse University, Syracuse, N. Y.*
"Analysis and Design of Tunnel Diode Oscillators," *R. T. Lynch.*
"The Tunnel Diode—Its Operation and Characteristics," *C. J. Neuse.*

Session III—Coherent Light

Chairman: *C. Shafer, Raytheon Co., Waltham, Mass.*
"Generation of Optical Harmonics," *P. Frankin, University of Michigan, Ann Arbor, Mich.*
"Further Investigations on Doped Calcium Fluoride Optical Masers," *P. Sorokin, IBM Research Center, Yorktown, N. Y.*
"Some Properties of a Gaseous Optical Maser," *C. F. Luck, R. A. Paananen, and H. Statz, Raytheon Co., Waltham, Mass.*
"Problems in an Optical Radar Design," *M. L. Stich, Hughes Aircraft Co., Malibu, Calif.*

Session IV—Information Theory

Chairman: *D. Van Meter, Litton Systems, Inc., Waltham, Mass.*
"The Approximation of Realizable Ambiguity Functions," *S. M. Sussman, Melpar, Inc., Watertown, Mass.*
"A System for Ordering Codes Based on Message Reliability," *R. D. Klein, Northeastern University, Boston, Mass.*
"Error Probabilities for *M*-ary Non-Orthogonal Signals Under Phase Coherent and Phase Incoherent Operation," *A. H. Nuttall, Litton Systems, Inc., Waltham, Mass.*
"Adaptive Redundancy Removal," *R. Van Blerkom, IBM Communications Center, Rockville, Md.*

Session V—Tunneling

Chairman: *J. M. Levine, Raytheon Co., Waltham, Mass.*
"New Effects in Superconductivity," *S. Shapiro, A. D. Little, Inc., Cambridge, Mass.*
"Experiments in Tunnel Emission," *C. A. Mead, California Inst. Tech., Pasadena, Calif.*
"Metal-Interface Amplifiers," *J. P. Spratt and R. F. Schwarz, Philco Corp., Blue Bell, Pa.*
"Tunneling in Esaki Diodes—A Survey Including New Effects," *W. Bernard, Raytheon Co., Waltham, Mass.*

Session VI—Reliability I—A Design Criterion

Chairman: *I. Bosinoff, Sylvania Electric Products, Inc., Waltham, Mass.*
"A Survey of Techniques for Analysis and Prediction of Equipment for Reliability," *H. E. Blanton, Raytheon Co., Lexington, Mass. and R. M. Jacobs, Sylvania Electric Products, Inc., Waltham, Mass.*
"Design Considerations for Reliable Electronic Equipment," *W. J. West, Autonetics, Downey, Calif.*
"Predicting the Reliability of Satellite Systems," *R. H. Myers, Hughes Aircraft Co., Culver City, Calif.*
"Reliability and Design of Electronics Systems and Equipment," *F. Moskowitz, Lab. for Electronics, Boston, Mass.*

Session VII—Computer Techniques

"A Low-Power, High-Speed Transistor-Tunnel-Diode Hybrid Logic Circuit," *J. A. Ekiss and P. M. Spiegel, Philco Corp., Lansdale, Pa.*
"Digital Circuits with a Minimum Number of Tunnel Diodes," *B. Rabinovici and J. Klapper, RCA, New York, N. Y.*
"A Thin Film Microcircuit Inductor," *T. Constantakes, General Instrument Corp., Hicksville, L. I., N. Y.*
"The Maximum Efficiency Concept for High-Speed Random-Access Memories," *M. M. Stern and H. A. Ullman, Sylvania Electric Products, Inc., Needham, Mass.*
"Diode and Transistor Logic in Synthesis of Symmetric Boolean Matrices," *H. K. Cooper, Pacific Semiconductors, Inc., Lawndale, Calif.*

Wednesday Morning, November 15

Session VIII—Radio Astronomy I

Chairman: *J. Ruze, Radiation Engrg. Lab., Maynard, Mass.*
"Fundamental Requirements of Radio Astronomical Measurements," *A. E. Lilley, Harvard College Observatory, Cambridge, Mass.*

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IR detector line has extreme sensitivity, fast response time, 1-30 micron coverage

Raytheon's rugged all-metal detectors are ideal for aerospace applications.

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Raytheon detectors provide coverage in the important long wavelength regions of 1 to 6, 6 to 9, 8 to 14, and 14 to 30 microns. Raytheon thin-film polarizer has 98% polarization at 4 microns and beyond.

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TYPE	MATERIAL	WINDOW	PACKAGE	RESPONSE
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QKN-1003	AuGe	BaF2	#1	1-9 microns
QKN-884	AuGe	Silicon	#2	1-9 microns
QKN-1004	AuGe	BaF2	#2	1-9 microns
QKN-961	CdGe	BaF2	#3	1-17 microns
QKN-902	CuGe	BaF2	#4	1-17 microns
QKN-1007	CdGe	KRS-5	#3	1-25 microns
QKN-1008	CdGe	KRS-5	#4	1-25 microns
QKN-1009	CuGe	KRS-5	#4	1-30 microns
IR Polarizer		98% polarization		4 microns & beyond

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MICROWAVE AND POWER TUBE DIVISION

"An Attempt to Detect the Galactic Deuterium Line Using Digital Autocorrelation Techniques," *S. Weinreb, Mass. Inst. Tech., Cambridge, Mass.*

"Radiometric Instrumentation at Eight Millimeters," *J. W. Meyer, J. J. Kirwan, and D. M. Towle, MIT Lincoln Lab., Lexington, Mass.*

"The $(1-\cos\phi)$ Radiometer," *J. W. McGinn, Jr., MIT Lincoln Lab., Lexington, Mass.*

Session IX—Solid-State Devices and Materials

Chairman: *H. G. Rudenberg, Transitron Electronic Corp., Boston, Mass.*

"Varactor Diodes for Microwave Applications," *K. E. Mortenson, Microwave Associates, Burlington, Mass.*

"Recent Advances in Transistors," *T. A. Longo, Sylvania Electric Products, Inc., Waltham, Mass.*

"Trends and Topics in New Materials," *W. C. Dunlap, Raytheon Co., Waltham, Mass.*

"The Performance of Solar Cells in Space Environments," *F. M. Smits, Bell Telephone Labs., Murray Hill, N. J.*

Session X—Reliability II—A Design Criterion

Chairman: *M. V. Ratynski, Electronic Systems Div., USAF, Hanscom Field, Bedford, Mass.*

"Design Reliability Creation Through Management Directives," *J. W. Griswold, The Boeing Co., Seattle, Wash.*

"The Human as an Element in System Reliability," *J. Spiegel, The Mitre Corp., Lexington, Mass.*

"Maintainability and Material Support Costs," *R. I. Moeller, IBM Space Guidance Center, Owego, N. Y.*

"Weapons System Design and Supportability—A Function of Failure Prediction," *V. J. Taylor, USAF Headquarters, Washington, D. C.*

Session XI—Superpower Microwave Tubes —A Panel Discussion on the Relative Merits of Linear-Beam and Cross-Field Superpower Microwave Tubes

Chairman: *J. Burgess, Rome Air Development Center, Rome N. Y.*

Panel Members: *T. D. Sege, Sperry Gyroscope Co., Great Neck, L. I., N. Y., and W. C. Brown, Raytheon Co., Waltham, Mass.*

Wednesday Afternoon

Session XII—Radio Astronomy II

Chairman: *P. Rizzi, Microwave Technology, Inc., Waltham, Mass.*

"A Broad-Band Interferometer for Studies of Discrete Radio Sources," *J. N. Douglas, Yale University, New Haven, Conn.*

"Mills Cross and Related Pencil-Beam Antennas," *G. Swarup, Stanford University, Stanford, Calif.*

"Electronically Scanned Radio Telescopes," *I. M. Faigen, Chu Associates, Harvard, Mass., and S. J. Goldstein, Jr., Harvard College Observatory, Cambridge, Mass.*

"The Design of Radio Astronomy Instruments," *H. G. Weiss, MIT Lincoln Lab., Lexington, Mass.*

"Report on the 1000-Foot Reflector Antenna," *P. Blacksmith and E. L. Bouche, Air Force Cambridge Research Center, Hanscom Field, Bedford, Mass.*

Session XIII—Microelectronics

Chairman: *D. E. Thomas, Sylvania Electric Products, Inc., Woburn, Mass.*

"Microwatt-Microelectronics," *W. Gaertner, CBS Labs., Stamford, Conn.*

"Scribe-plating and Trace-plating; New Techniques for Microelectronic Semiconductors," *W. Rindner and J. M. Levine, Raytheon Co., Waltham, Mass.*

"Titanium Thin-Film Circuits," *W. C. Fuller, Lockheed Aircraft Corp., Sunnyvale, Calif.*

"Solid-State Micrologic Elements," *N. Gault, L. Kattner, and J. Nall, Fairchild Semiconductor Corp., Mountain View, Calif.*

Thursday Morning, November 16

Session XIV—Bio-Medical Electronics I —Machine Deciphering of Biological Data

Chairman: *S. Aronow, Massachusetts General Hospital, Boston, Mass.*

"Electrical Signals and Noise in Bio-Medical Research," *F. R. Ervin, Massachusetts General Hospital, Boston, Mass.*

"Theoretical Basis for Biological Signal Generation," *J. Cowan, MIT, Cambridge, Mass.*

"A Basic Digital System for Biological Research," *S. F. Molner, Epsco, Inc., Cambridge, Mass.*

Display and Computing of Biological Data," *T. A. Kalin, Air Force Cambridge Res. Lab., Bedford, Mass.*

"Bioelectrical Recording—General Requirements and a Specific Example," *B. S. Rosner, West Haven Veterans Administration Hospital, West Haven, Conn.*

Session XV—Atomic Frequency and Time Standards

Chairman: *P. Bender, National Bureau of Standards, Washington, D. C.*

"The Cesium Atomic Beam Frequency Standard," *J. Holloway, Bomac Labs., Inc., Beverly, Mass.*

"The Rubidium Optically-Pumped Frequency Standard," *R. Rempel, Varian Associates, Palo Alto, Calif.*

"The Hydrogen Maser," *D. Kleppner, Harvard University, Cambridge, Mass.*

"Stable Radio Frequencies for Atomic Standards," *C. Searle, MIT, Cambridge, Mass.*

Session XVI—University-Industry Cooperative Programs that Enable Engineers to Keep Abreast of Latest Technological Developments—A Panel Discussion

Chairman: *R. M. Fano, MIT, Cambridge, Mass.*

Panel Members: *J. Thomas, Minneapolis-Honeywell Regulator Co., Boston, Mass.; N. Marcuvitz, Polytechnic Inst., Brooklyn, Brooklyn, N. Y.; R. Villers, Columbia University, and Rautenstrouch and Villers, New York, N. Y., and J. Wolfe, General Electric Co., Schenactady, N. Y.*

Session XVII—Microwave Solid-State Devices I

Chairman: *G. Wade, Raytheon Co., Burlington, Mass.*

"Generation of Coherent Light," *H. Statz, Raytheon Co., Waltham, Mass.*

"Microwave Modulation of Light by the Electrooptic Effect," *I. P. Kaminow, Bell Telephone Labs., Holmdel, N. J.*

"Microwave Demodulation of Light," *A. E. Siegman, Stanford University, Stanford, Calif.*

"Parametric Amplifiers—A Status Report," *R. Adler, Zenith Radio Corp., Chicago, Ill.*

Thursday Afternoon

Session XVIII—Bio-Medical Electronics II—Human Performance and Design

Chairman: *A. T. Kornfield, Biosearch Co., Boston, Mass.*

"Life Support by Control of 'External Environment' in Closed Compartments," *A. Bialecki, Electric Boat Co., Groton, Conn.*

"Life Support by Control of 'Internal Environment'—Artificial Hearts, Lungs and Pacemakers," *A. Kantrowitz, Maimonides Hospital, Brooklyn, N. Y.*

"Sensory Aids—Implants and Endosondes," *V. K. Zworykin, RCA Labs., Princeton, N. J.*

"Action Aids—Artificial Limbs and Other Prosthetic Devices," *E. Murphy, Prosthetics Research Center, Veterans Administration, New York, N. Y.*

Session XIX—Atomic Frequency and Time Standards II

Chairman: *J. King, MIT, Cambridge, Mass.*

"Microwave Phase-Locking Systems," *J. L. Altman, Lab. for Electronics, Boston, Mass.*

"World-Wide Clock Synchronization," *F. H. Reder, U. S. Army Signal Research and Dev. Lab., Ft. Monmouth, N. J.*

"Time Standard Requirements in Satellite Tracking for Navigation and Geodesy," *G. Weiffenbach, Johns Hopkins University, Baltimore, Md.*

"On the General Use of Atomichrons," *E. Grant, National Co., Malden, Mass.*

"An Atomic Definition for Unit of Time," *J. M. Richardson, Boulder Labs., Natl. Bureau of Standards, Boulder, Colo.*

"Time Jitter and Its Measurement," *R. G. Heaton, Lab. for Electronics, Boston, Mass.*

Session XX—Dynamics of Research and Development—A Panel Discussion

Chairman: *R. L. Blanchard, Trans-Sonics, Inc., Lexington, Mass.*

Panel Members: *W. K. Linvill, Stanford University, Stanford, Calif.; L. I. Beranek, Bolt, Beranek and Newman, Inc., Cambridge, Mass.; J. W. Forrester, MIT, Cambridge, Mass.; O. T. Fundingsland, Raytheon Co., Lexington, Mass.; and J. E. Storer, Sylvania Electronic Products, Inc., Waltham, Mass.*

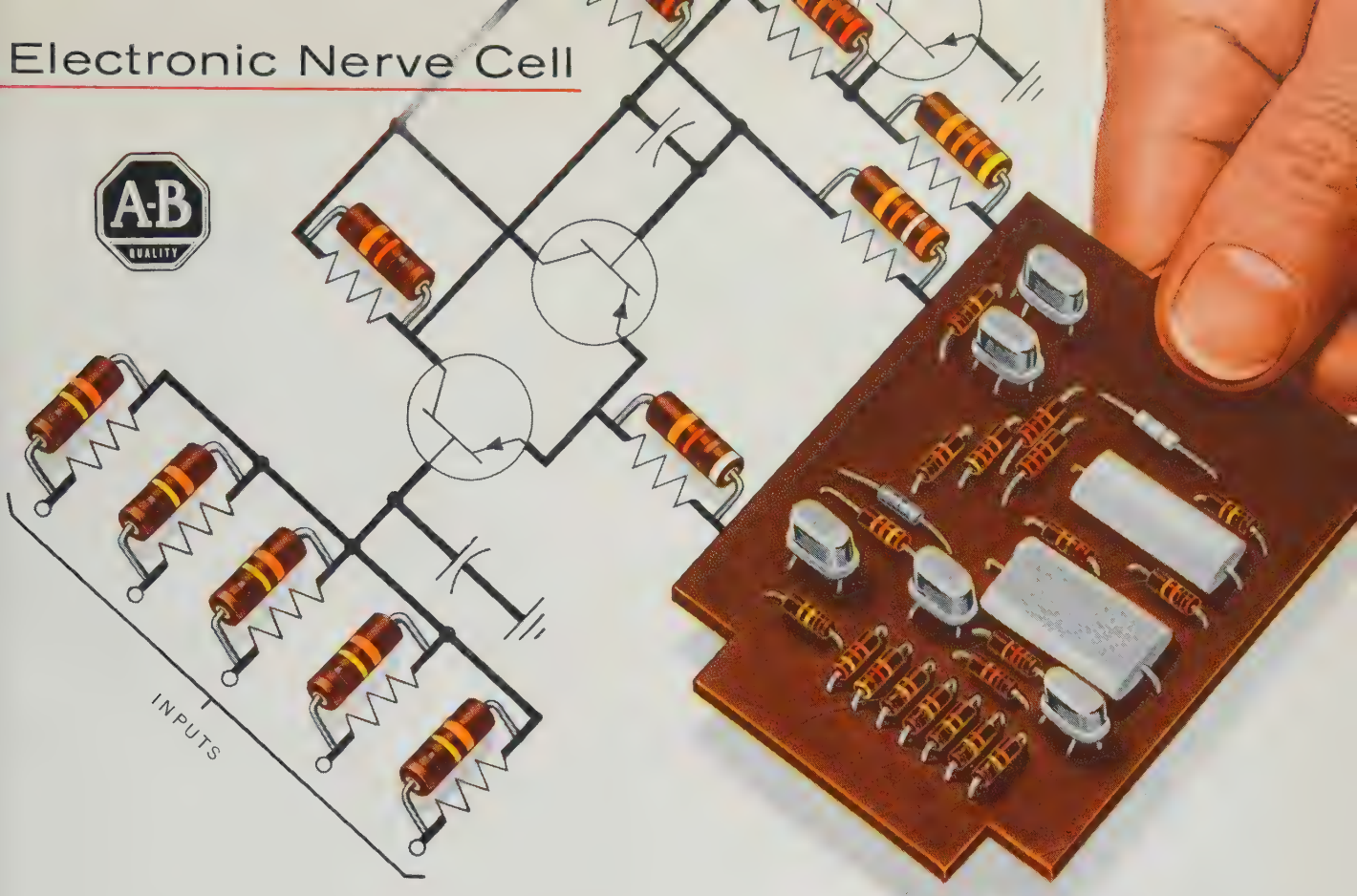
Session XXI—Microwave Solid-State Devices II

"Semiconductor Microwave Control Devices," *L. Gould, M. Bloom, and R. Tennenholtz, Microwave Associates, Inc., Burlington, Mass.*

"Solid-State Microwave Amplifiers," *J. Sie, Microstate, Inc., Murray Hill, N. J.*

"Varactor Harmonic Generators," *M. E. Hines, Microwave Associates, Inc., Burlington, Mass.*

Electronic Nerve Cell



Illustrated above is a section of the schematic diagram for the artificial neuron (nerve cell).

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Tuning breakage, backlash, and accelerated wear are among the problems you encounter in a trimmer capacitor whose core *rotates* during tuning.

What is a direct traverse trimmer and why?

That's why we took the rotation out of our trimmer capacitors. Our core runs up and down its tube without turning.

That's why direct traverse tuning curves are all smooth lines, utterly devoid of capacitance reversals.

That's why direct traverse trimmers tune so smoothly, without a snag to cause breakage just when you think the circuit is complete and ready to go.

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Plus the properties of glass. We've added to this direct traverse design the many values of glass. No other material combines such high reliability with such low TC. Or such precision at such low cost. Let the specs speak for themselves:

TC	±50 to ±100
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Megohms, IR	10^6
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Four models. Where space is no problem, you'll look immediately for our standard direct traverse trimmers. They range from .5-3.0 to 1-12 uuf. Approximately 0.6 uuf change per turn.

When space is tight both in front of and behind your panel, you'll appreciate our petite mini-trimmers. Not only are the over-all dimensions small, but we throw in fixed cavity tuning which keeps the screw enclosed at all times. These range from 1-4.5 to 1-18 uuf with approximately 0.40 uuf change per turn.

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7.3.2 Gas-Filled Radiation Counter Tubes

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7.4 Camera Tubes, Phototubes, and Storage Tubes in Which Photo-Emission is Essential

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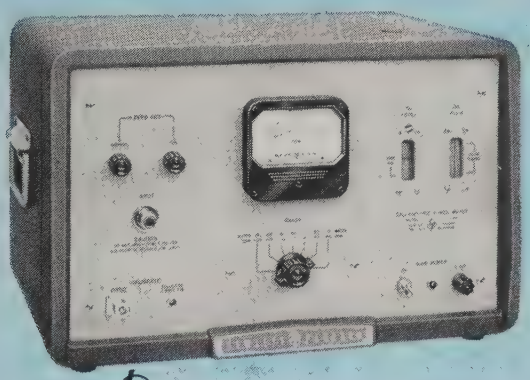

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Just connect, then read power
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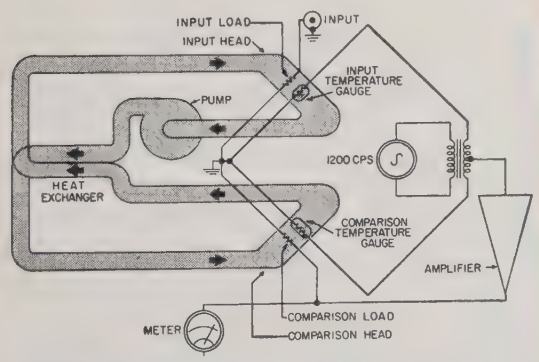
hp 434A makes it just this simple: connect your power source to the front-panel input terminal . . . then read power directly in watts and DBW, 7 ranges, full-scale readings 10 mw to 10 watts average, 1 KW peak!

The hp 434A is a self-contained, high-precision instrument, yet it operates with only two simple front-panel controls—range switch and zero set. No other adjustments are needed to measure power at any frequency, dc to 12.4 GC. And the compact 434A requires no thermistor, barretter, external terminations or plumbing of any kind! Thus it is ideally suited for use by non-technical personnel to measure AM power, pulsed power, cw, video and dc power. An internal calibrator ensures that the 434A operates always at peak performance.

Fast response—high stability

hp 434A provides a full scale response time of 5 seconds or less, by means of a self-balancing bridge and a high-efficiency heat transfer system using an oil stream (see diagram). This fast reaction, a fraction of the response time needed by ordinary calorimeters, means the 434A quickly follows small changes in input tuning circuits. Further, the 434A achieves a new high in stability through the use of twin power-sensitive elements in a single oil stream, a design which makes the accuracy independent of changes in oil flow rate and temperature.



BRIEF SPECIFICATIONS

Input Power Range: 7 ranges; full-scale readings 0.01 to 10 watts, with continuous readings —30 to 10 DBW; may be extended upward with attenuators or directional couplers. Peak input power: 1 KW maximum.

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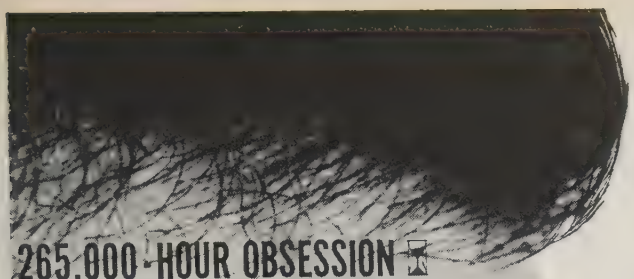

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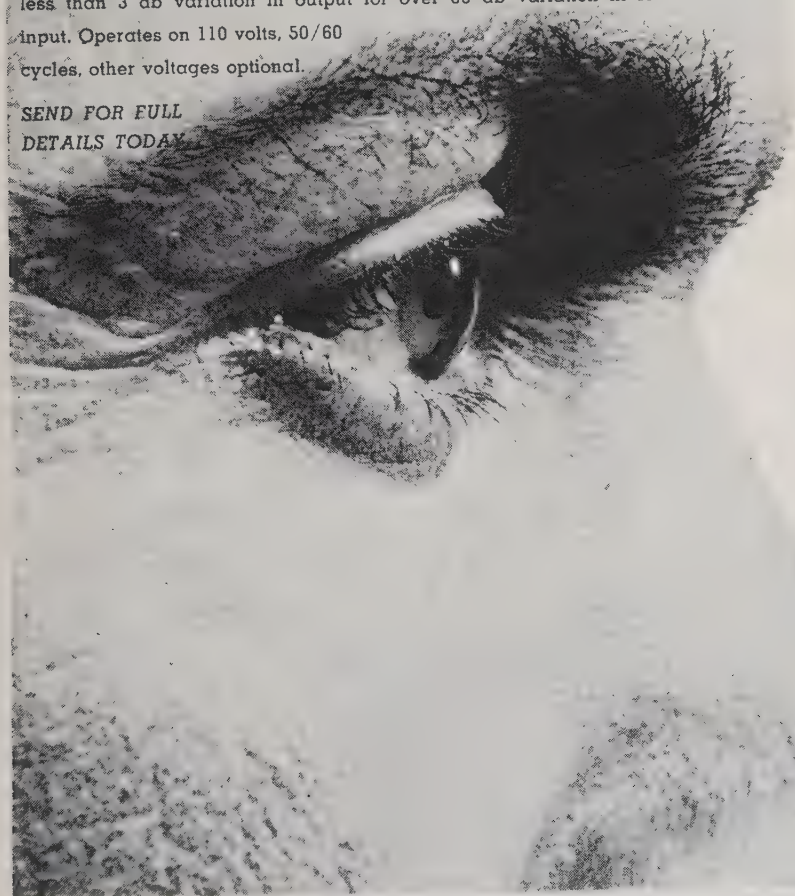

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28.4.3 Definitions and Letter Symbols of Semiconductors

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21.7 Letter Symbols

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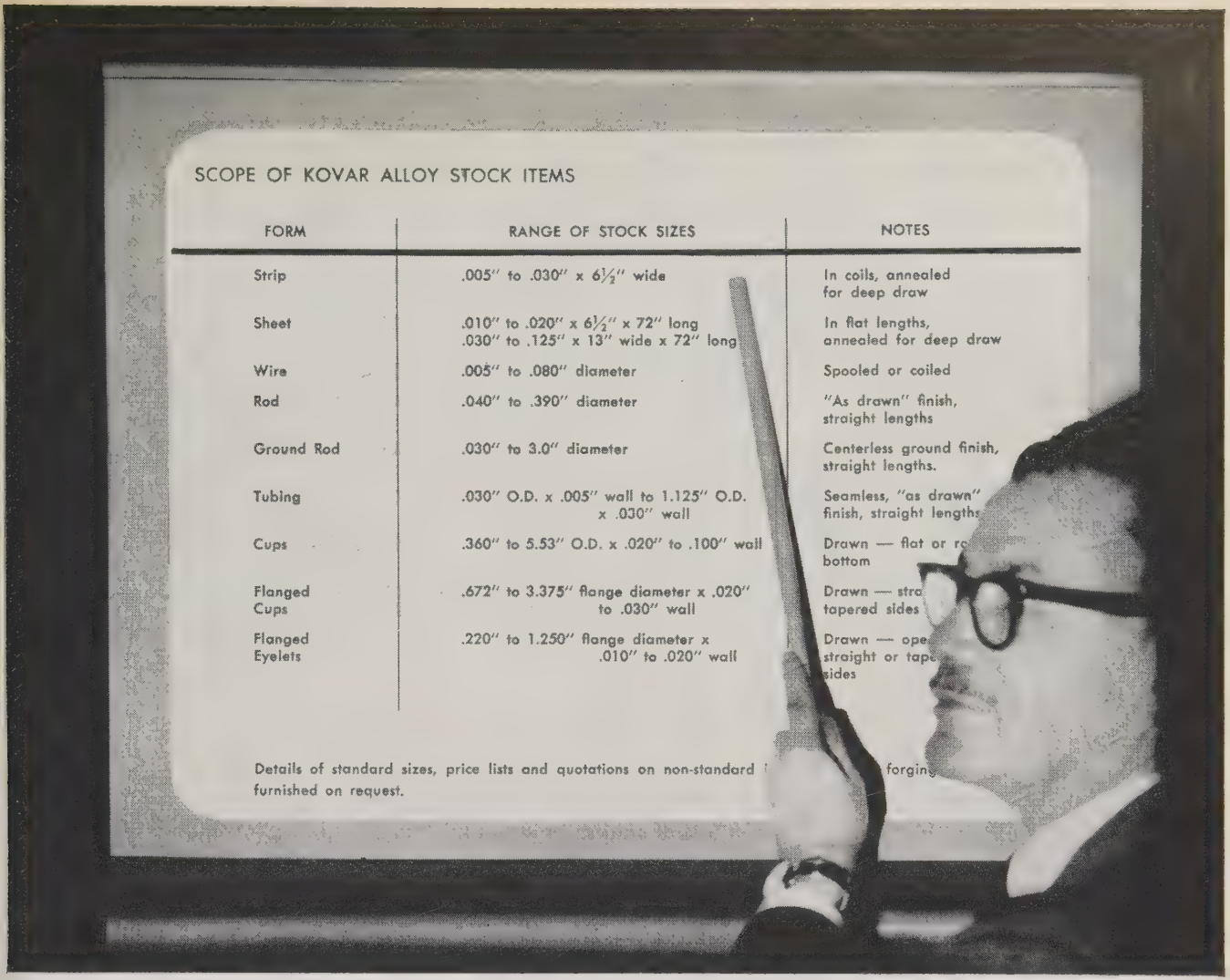
SCOPE OF KOVAR ALLOY STOCK ITEMS
FORM
RANGE OF STOCK SIZES
NOTES
Strip
.005" to .030" x 6½" wide
In coils, annealed for deep draw
Sheet
.010" to .020" x 6½" x 72" long
.030" to .125" x 13" wide x 72" long
In flat lengths, annealed for deep draw
Wire
.005" to .080" diameter
Spooled or coiled
Rod
.040" to .390" diameter
"As drawn" finish, straight lengths
Ground Rod
.030" to 3.0" diameter
Centerless ground finish, straight lengths.
Tubing
.030" O.D. x .005" wall to 1.125" O.D. x .030" wall
Seamless, "as drawn" finish, straight lengths
Cups
.360" to 5.53" O.D. x .020" to .100" wall
Drawn — flat or r bottom
Flanged Cups
.672" to 3.375" flange diameter x .020" to .030" wall
Drawn — str tapered sides
Flanged Eyelets
.220" to 1.250" flange diameter x .010" to .020" wall
Drawn — ope straight or tap sides
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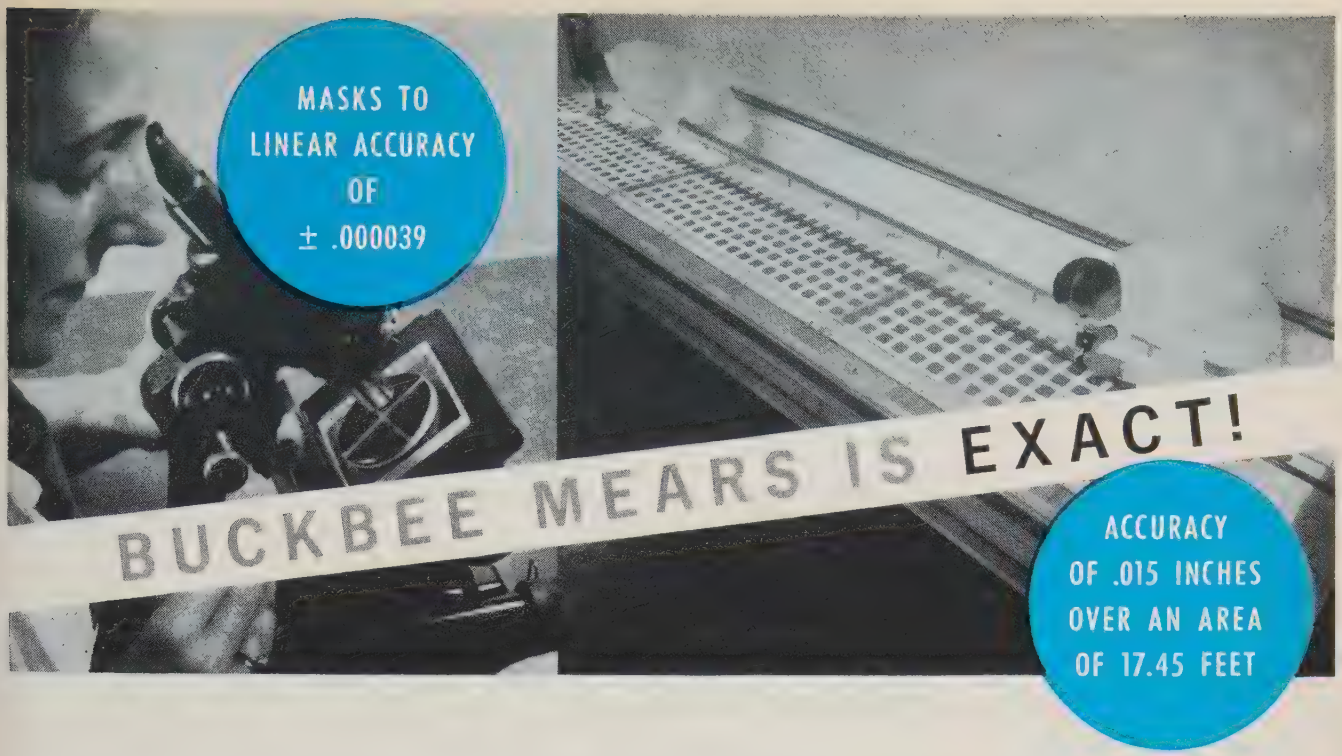

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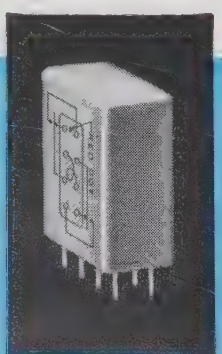
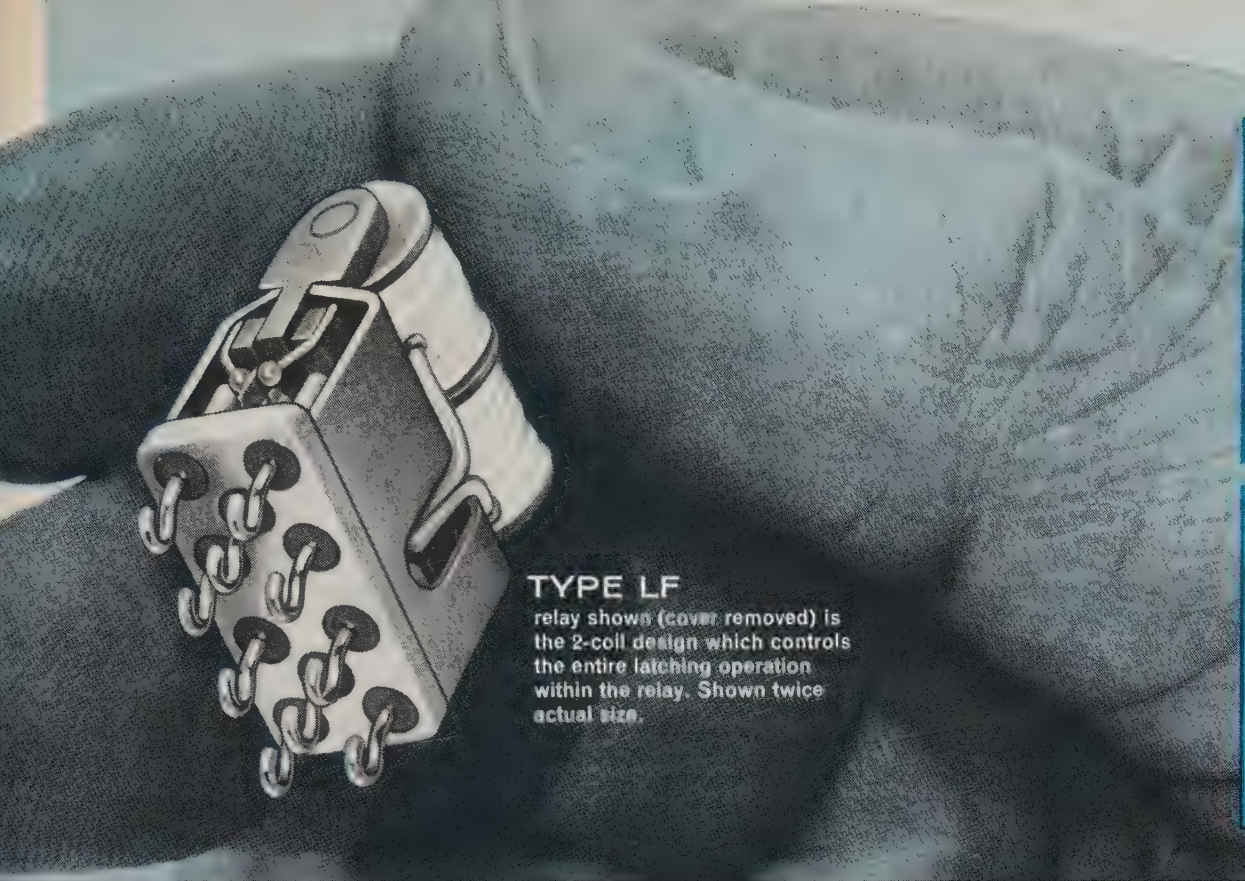
The Type LF is available with either 2-coil or 1-coil configuration. The 2-coil relay allows complete control of the latching operation within the relay and provides an extremely compact operating unit. The 1-coil relay is somewhat more sensitive; it is adaptable to existing circuits where outside control is provided. (See opposite page for specifications and circuit diagrams.) The Type LF provides the same wide range of mounting arrangements and terminals as the CLARE Type F relay.

FOR NON-LATCHING OPERATION



CLARE Type F Subminiature Crystal Can Relay

The CLARE Type F relay is extremely fast and more than moderately sensitive. It is built to withstand temperature extremes, heavy shock and extreme vibration. Contacts, rated at 3 amperes, are excellent for low-level circuit operations. Send for Design Manual 203.



TYPE LF
relay shown (cover removed) is the 2-coil design which controls the entire latching operation within the relay. Shown twice actual size.

2-Coil Circuit Diagram

1-Coil Circuit Diagram

PHYSICAL FEATURES

Life Expectancy
Wet Circuit:
3.0 amperes, 28VDC resistive—100,000 operations
2.0 amperes, 28VDC resistive—250,000 operations
1.0 ampere, 28VDC resistive—1,000,000 operations
1.0 ampere, 28VDC inductive (100 millihenry)—100,000 operations
1.0 ampere, 115 VAC resistive—100,000 operations
Dry Circuit:
1,000,000 miss free operations when subject to conventional dry circuit requirements.

Temperature—+125° C to −65° C

Shock—100g's for 1/2 sine wave 11 ± 1 MS pulse

Linear Acceleration—100g's minimum

Vibration—.250″ DA or 30 g's, 5-2000 cps.

Humidity & Salt Spray—MIL-R-5757D

Enclosures: Tinned brass cover with fungus-resistant finish. Hermetically sealed and filled with dry nitrogen at atmospheric pressure.

Contact Arrangement—2PDT latching

Terminals—Plug-in (3/16″ straight), solder hook, 3″ straight

Wiring—Two coils (as shown on drawing above)
One coil (as shown on drawing above)

Weights—.54 oz. for plug-in
.62 oz. for 2 studs, 3″ leads

ELECTRICAL FEATURES

Operate Time—Two coil: When applying—for a minimum of 5 milliseconds—a voltage of at least two times the must operate voltage, the operate time including bounce will not exceed 5 milliseconds. One Coil: operate time will not exceed 8 milliseconds.

Sensitivity—Two coil, approximately 150 milliwatts
One coil, approximately 75 milliwatts

Dielectric Strength
Sea level: 1000 volts rms—all terminals to case
1000 volts rms—between contact sets
600 volts rms—between open contacts of a set
70,000 ft: 350 volts rms—all terminals to case

Insulation Resistance—1000 megohms minimum at +125° C between any two terminals and between all terminals and case.

Maximum Interelectrode Capacitance—

Closed contacts to case	3.7 picofarads
Open contacts to case	2.0 picofarads
Between contacts of a set	2.0 picofarads
Between adjacent contact sets	3.5 picofarads

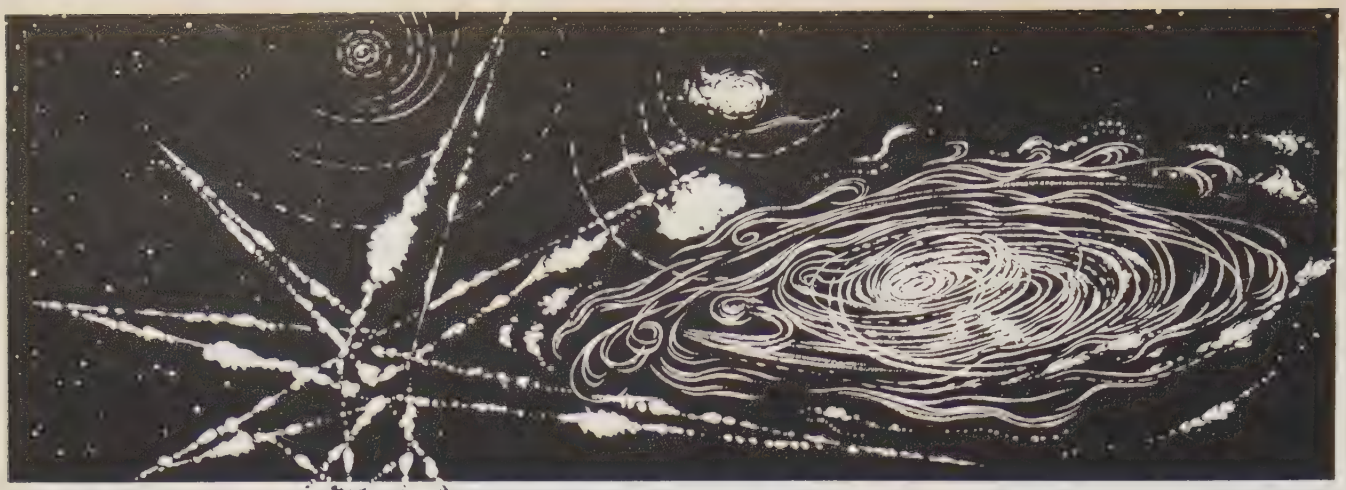
Maximum Coil Dissipation
Two Coil: .50 watts at +125° C
.75 watts at +25° C
One Coil: 1.25 watts at +125° C
2.0 watts at +25° C

Standard Adjustment—Relay will operate and hold when the must operate voltage is applied

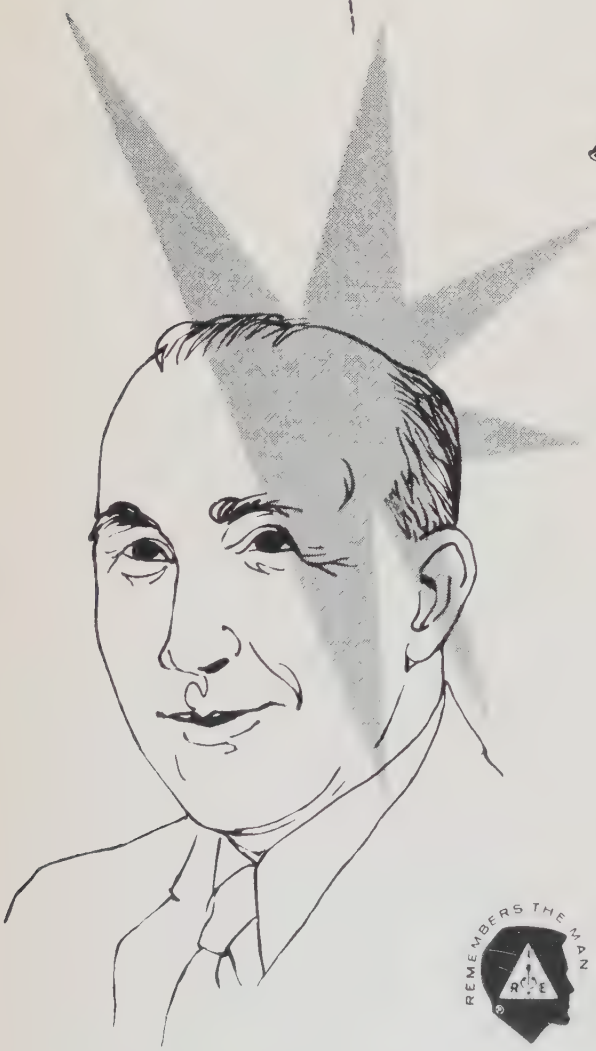
Contact Resistance:
Maximum: 50 milliohms at 6 volts, 100 milliamperes.
Typical: 25 milliohms at 6 volts, 100 milliamperes.

For coil and mounting data on CLARE Type LF relay send for CPC-12. Address: C. P. Clare & Co., 3101 Pratt Blvd., Chicago 45, Illinois. In Canada: C. P. Clare Canada Ltd., 840 Caledonia Road, Toronto 19, Ontario. Cable Address: CLARELAY.

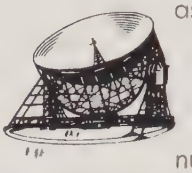
C. P. CLARE & CO. *Relays and related control components*



WHY STAR GAZE?



From Alaska to Australia and all around the world, scientists scan the sky with giant radio-telescopes, "listening" for the music of the spheres. To these radio astronomers the night is never silent. Age-less unseen stars talk to them; and from the mass of hisses and crackles, elec-tronic instruments pinpoint a growing number of new friends, making up a radio map of the sky far different from the sky as we know it. These dedicated scientists include hundreds of members of The Institute of Radio Engineers. Through them and their writings in *Proceedings of the IRE*, a world-wide pool of knowledge is brought to our nation.



While some IRE members are thus reaching out beyond our Universe, many thousands of others are working as devotedly on matters that concern our own planet. The IRE encourages both, rewarding research that is useful but not necessarily spectacular.

Fittingly, the IRE's Medal of Honor, highest annual tech-nical award in the field of electronics, goes in 1961 to a man whose researches apply not just to our own world but to possible links with other worlds as well. His field: communications networks. Dr. Ernst A. Guillemin gets the award for "Outstanding scientific and engineering achievements."

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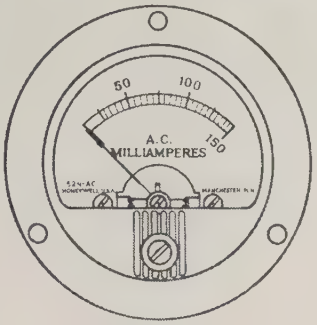
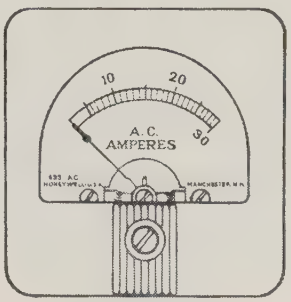
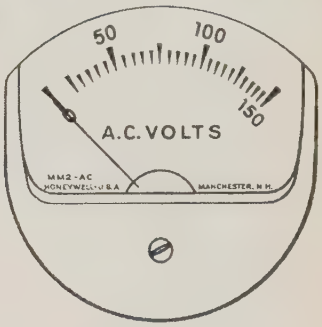


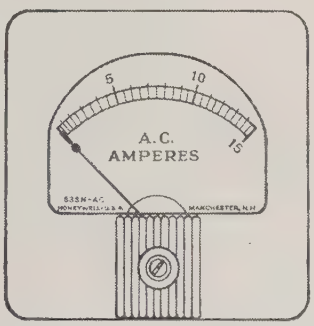
52N-AC 2 1/2"



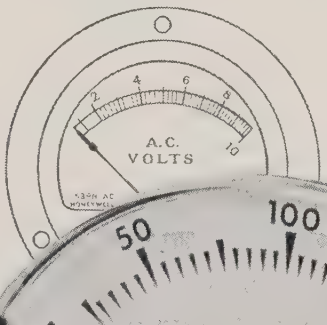
52S-AC 2 1/2"

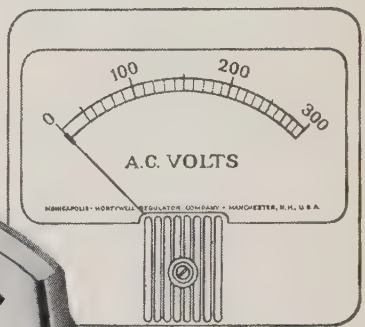


MM2-AC 2 1/2"

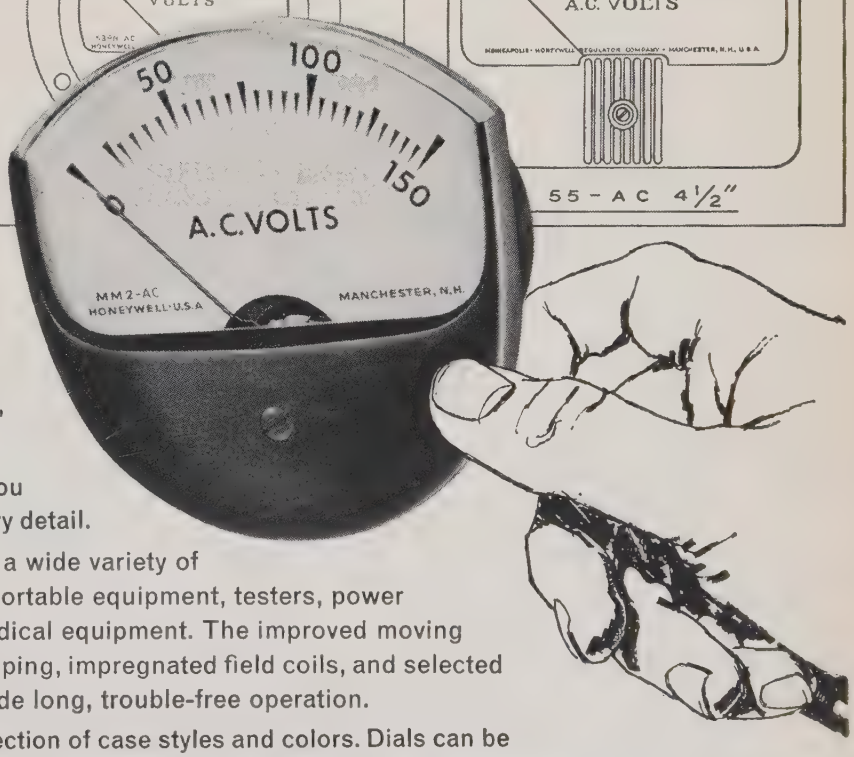


53SN-AC 3 1/2"





55-AC 4 1/2"



Here are the AC counterparts of Honeywell's popular DC panel meters. Iron Vane AC Meters are perfectly matched to the DC range and are available in both the Medalist and "standard" case styles. This means a minimum of trouble and expense in mounting. And you are assured of harmonious styling in every detail.

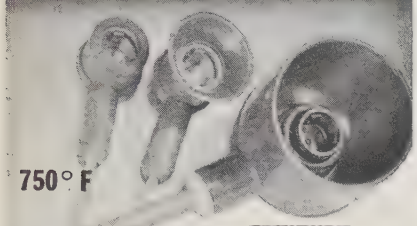
Iron Vane AC Meters are designed for a wide variety of commercial applications — including portable equipment, testers, power supplies, generator equipment and medical equipment. The improved moving iron mechanism features magnetic damping, impregnated field coils, and selected fixed and moving iron material to provide long, trouble-free operation.

These meters are available in a wide selection of case styles and colors. Dials can be custom designed with your company name, trade-mark or other data. For full information, contact our representative in your area — he's listed in your classified telephone directory. Or us: Precision Meter Division, Minneapolis-Honeywell Regulator Co., Manchester, N.H., U.S.A. In Canada, Honeywell Controls Limited, Toronto 17, Ontario and around the world: HONEYWELL INTERNATIONAL — Sales and service offices in all principal cities of the world.

Honeywell

H HONEYWELL Precision Meters

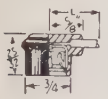
SILICONE INSULATED TUBE CAP CONNECTORS



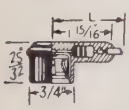
HI ALTITUDE TEMPERATURE VOLTAGE

This is a new series of Tube Cap Connectors using special silicone components for high reliability applications. They provide the highest degree of resistance to temperature extremes and are virtually unaffected by ozone and corona. The excellent dielectric characteristics make them ideal for high voltage. Skirts and sealed-in leads guard against flashover at high altitudes. Additional features include anti-corona cup and long-life spring contacts.

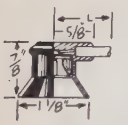
Clip this out — keep handy for part numbers and specs on connectors below for either ¼" or ⅜" top caps. Prefix 90 for ¼"; 91 for ⅜". Lead wire 18" long from center of cap or length to your specs.



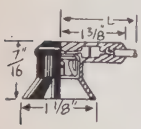
#90 or 91SCCSL beryllium copper contact, cadmium plated nests in anti-corona cup. Silicone rubber insulation throughout.



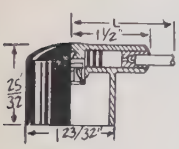
#90 or 91SCCRSL beryllium copper contact, cadmium plated nests in anti-corona cup. Silicone rubber insulation throughout. Takes up to one watt resistor — specify value and tolerance.



#90 or 91SCCDSL beryllium copper contact, cadmium plated nests in anti-corona cup. Skirt clings to tube — guards against flash-over Silicone rubber insulation throughout.



#90 or 91SCCDRSL beryllium copper contact, cadmium plated enclosed in anti-corona cup. Skirt clings to tube — helps suppress corona—guards against arc-over. Takes up to one watt resistor. Specify value and tolerance.



#90 or 91CCSTLRL beryllium copper contact, cadmium plated nests in anti-corona cup. Glass-filled silicone insulation on cap; silicone rubber on lead. Long skirt for arc-over. Takes up to 2 watt resistor. Specify value and tolerance.

Besides new silicone types — Alden provides a complete series of connectors for ¼", ⅜" and 9/16" cap in your choice of phenolic, mica, polyethylene, nylon and Kel-F. Complete hi-voltage cable assembles are available using Alden hi-voltage disconnects and tube cap connectors.

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IRE People

Alan D. Bailey (M'58) has joined the staff as Senior Electronic Engineer in the Space Instrumentation Division of Technology Instrument Corporation, Acton, Mass.

A. D. Bailey

He is responsible for electrical and electronic instrumentation flown on research sounding rockets, deep space probes and satellite vehicles.

Formerly, he was an Assistant Professor at Northeastern University, Boston, Mass., in electronic research and provided technical supervision of the design and construction of instrumentation for rocket-borne packages for atmospheric geophysical measurements. Previously, he was with Anderson Nichols, Inc., engineering electronic circuit design for gating and timing and evaluated small digital computers. He also did circuit design through Northeastern in support of a theoretical group under contract with the Air Force Cambridge Research Center.

Mr. Bailey holds the Master's degree in electrical engineering from Northeastern University. He is a member of the American Institute of Electronic Engineers and the American Rocket Society.

❖

Jack W. Blanchard (M'57) has been appointed chief engineer of the Phoenix Electronics Plant of Kaiser Aircraft & Electronics Division of Kaiser Industries Corporation. He will direct the Kaiser-Phoenix Electronics Plant's engineering activities in the design of electronic systems and equipment under development for aircraft and missile checkout and ground support.

J. W. Blanchard

Prior to joining Kaiser, he was Senior Project Director and Associate Research Engineer with the University of Arizona in Tucson, Arizona.

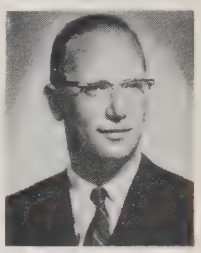
❖

G. H. Bateman (A'47–SM'51) has joined the Advanced Systems Department of Electro-Mechanical Research, Inc. as an engineer.

Mr. Bateman worked for Lear, Inc. of Grand Rapids, Michigan, as a Staff Consultant for electronics. Following this he had a private consulting business in the product design area. Prior to joining EMR, he was employed at The Martin Company, Orlando, Florida, as a member of their research and technical staff. In the Advanced Systems Department, he will be concentrating on research and development in the area of optical communications systems. Mr. Bateman received his BA degree from Boston University, Boston, Mass. He later studied electrical engineering at Northwestern University, Michigan State University and Massachusetts Institute of Technology.

❖

The appointment of **Dr. Guy Black** (A'55) as staff engineering specialist in development planning for Sylvania Electronic Systems, a division of Sylvania Electric Products Inc., has been announced recently. In his new capacity, Dr. Black will study trends in advanced electronic technology and will make recommendations concerning Sylvania Electronic Systems' planning for technological requirements of the future. He joined Sylvania in 1955 as a senior engineer at Sylvania's Electronic Defense Laboratories, Mountain View, Calif., where he performed operations reserarch, logistics problem analysis, and reliability and maintenance studies. In 1959, he was transferred to Waltham as a divisional staff analyst studying industrial applications of electronics.



G. Black

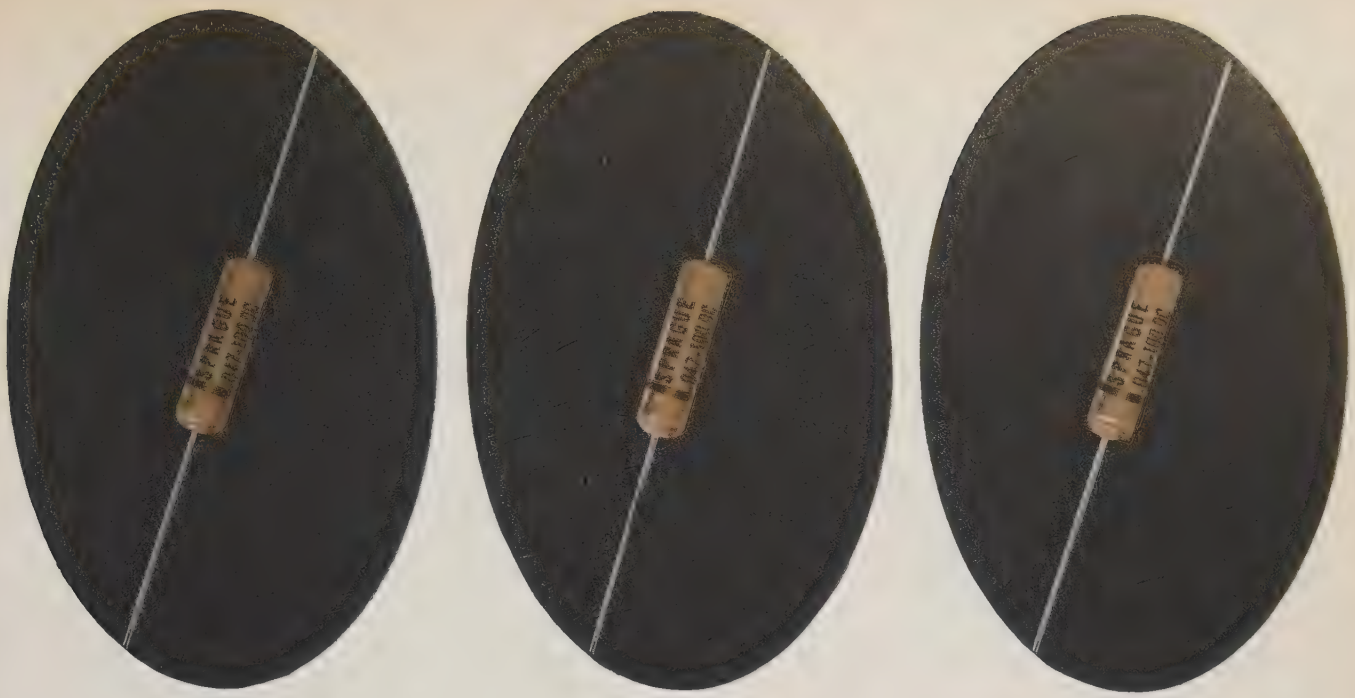
Prior to joining Sylvania, he was an assistant professor at the University of California for seven years. During this period, he taught economics and finance, and performed research studies on marketing, distribution and transportation of consumer food products. He has also served as an economist with the U. S. Department of Agriculture, studying trends in food consumption.

A graduate of Harvard University in 1941, Dr. Black later received his master's and doctorate in economics from the University of Chicago. He is the author of numerous articles on various aspects of marketing. He is a member of the Scientific Research Society of America, American Economic Association, and National Planning Association.

❖

Hugh S. Christian (S'46–A'48–M'55) has been employed by Electron Corpora-

(Continued on page 58A)



TINY...

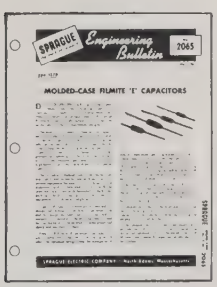
Latest space-maker for size-conscious designers of transistorized commercial and entertainment equipment is the new *Sprague Type 157P Molded-case Filmite® "E" Capacitor,* which combines unusual compactness with exceptional performance characteristics.

TAN...

Distinctive *tan coloring* identifies the Type 157P Capacitor and serves as your warranty of outstanding shock-and-humidity resistance. The tough molded armor also protects against possible damage during soldering operations, or changes in capacitance from mechanical pressure where wrapped capacitors are clamped or cast in assemblies.

TERRIFIC!

Standard operating temperature range is —40 C to +85 C. And with voltage derating, this outstanding capacitor may be operated to +105 C! Its high insulation resistance (due to the polyester film dielectric and molded housing) is another characteristic which qualifies the 157P Capacitor for critical coupling applications.



For complete technical data on Type 157P Filmite "E" Capacitors, write for Engineering Bulletin 2065 to Technical Literature Section, Sprague Electric Company, 235 Marshall Street, North Adams, Mass.

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TRANSISTORS	PULSE-FORMING NETWORKS	FUNCTIONAL DIGITAL CIRCUITS

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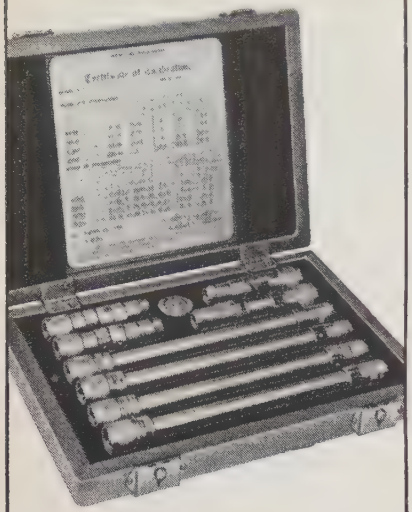

MODEL AS-1

PRECISION ATTENUATOR SET

This set contains eight Weinschel precision coaxial attenuators in a solid walnut, velvet-lined case. Included in the set are Weinschel Model 210 attenuators and Weinschel Model 50 attenuators, in attenuation values of 3, 6, 10, and 20 db. The set covers the frequency range of DC to 12.4 kmc. All eight attenuators are made with Weinschel's own stable film resistors, and have stainless steel bodies and stainless steel connectors which give maximum life with minimum wear.

CERTIFICATE OF CALIBRATION

A Certificate of Calibration showing calibration data for each attenuator is supplied with the set. The certificate gives DC resistance and insertion loss at three frequencies for each attenuator.

WEINSCHEL ENGINEERING
10501 METROPOLITAN AVENUE
KENSINGTON, MARYLAND

Certificate of Calibration

MODEL AS-1 SERIAL NO. 1

Insertion Loss in 50 Ohm System at 20°C ±3°C, 1 Milliwatt CW Input Power Maximum

MODEL 210 ATTENUATORS

Serial Number	Nominal Value	DC Resistance in Ohms	Insertion Loss Source and Load VSWR			
			< 1.03	< 1.06	< 1.06	< 1.10
		Female To Male	10 KMC	20 KMC	40 KMC	100 KMC
20059	3 db	34.46	3.00 ± .04	X	3.1 ± .15	3.2 ± .10
24469	6 db	67.70	5.65 ± .05	X	6.0 ± .15	6.0 ± .15
23547	10 db	113.60	9.05 ± .08	X	10.0 ± .17	10.1 ± .20
23408	20 db	228.40	X	18.5 ± .35	19.8 ± .27	20.6 ± .33

MODEL 50 ATTENUATORS

Serial Number	Nominal Value	DC Resistance In Ohms			Insertion Loss			
		Female to Male	Female to Ground	Male to Ground	DC	Source and Load VSWR < 1.03		
						4 KMC	10 KMC	
22189	3 db	17.15	151.10	151.10	3.01 ± .01	3.05 ± .03	3.10 ± .03	
22610	6 db	32.69	81.35	81.35	6.03 ± .01	6.10 ± .03	6.10 ± .03	
22797	10 db	51.92	81.04	81.00	10.01 ± .01	10.05 ± .03	10.00 ± .03	
24310	20 db	82.20	51.26	51.16	20.02 ± .01	20.10 ± .05	20.10 ± .06	

Date August 15, 1960 Test Engineer

Weinschel Eng. Job No. 0-4566 Chief Engineer

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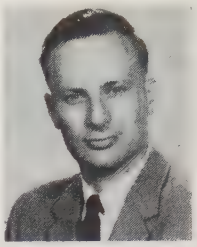
IRE People

(Continued from page 56A)

tion, subsidiary of Ling-Temco Electronics, Inc., as Administrative Engineering Assistant to President M. Zimmerman.



H. S. CHRISTIAN

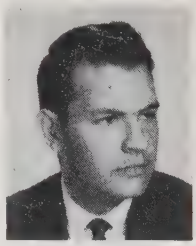
He was formerly Chief Engineer and Assistant to the General Manager of Diamond Power Specialty Corporation, a division of Babcock & Wilcox. He also has been Chief Engineer of the Microwave Division of Motorola, Inc., of Chicago, Ill., and Color Television Engineer for Raytheon Manufacturing Corporation of Chicago.

Mr. Christian is a graduate of the Illinois Institute of Technology, Chicago, and is a member of a number of professional societies.

Electron specializes in the production of low-cost television systems for commercial broadcasting, educational television, and commercial and industrial closed-circuit television as well as low-cost automated FM radio broadcasting systems.

❖

James R. Diamond (S'57–M'58) has been promoted to the position of Section Chief, Instrumentation Section with the Eastern Ground Electronics Engineering and Installation Agency (EGEEIA) located at Brookley AFB, Mobile, Ala. He graduated from Auburn University in June, 1957, and holds the Bachelor of Science Degree in electrical engineering. He is presently Chairman of the Publicity Committee of the Mobile Section of the IRE.



J. R. DIAMOND

He has been with EGEEIA since January, 1959, and has had project engineer responsibilities for design and contract monitoring of the Eglin Gulf Test Range Communications System. In his new position he will direct and supervise engineering of new and augmentation requirements of the EGTR Communications System and other microwave and voice communications facilities of the Air Proving Ground Center, Eglin AFB, Fla.

❖

Dr. Morris Ettenberg (A'45–M'50–SM'51), research professor of electrophysics at the Polytechnic Institute of Brooklyn, has accepted a one-year visiting professorship at the Technion-Israel Institute of Technology, Haifa, Israel. He will travel to Israel in an offbeat fashion for a visiting educator. Sailing to Sweden along with his car, he will ferry to Helsinki, Finland; drive through Finland and Russia (one month) to Odessa on the Black Sea; board another boat to Athens, Greece, and finally to Haifa. He expects to arrive in Israel on October 6 and begin teaching at the Technion on October 15. His wife Sylvia, daughter Israela, age 12, and son David, age 6, will join him later in the year.

Dr. Ettenberg received the invitation to be a visiting Fulbright professor of electrical engineering some years ago, but his work kept him from accepting until now. While there, he will also do research in the general area of microwaves and plasmas.

Dr. Ettenberg received his bachelor's degree in mathematics from City College in 1935, and his master's the following year also from City. In 1949, he received his Ph.D. from New York University in physics.

❖

Dr. Lester M. Field (M'48–F'52) has been named Assistant Group Executive of Hughes Aircraft Company's Components Group.

L. M. FIELD

In his new position, he will assist in directing various phases of the group's operations and coordinate further broadening of its expanding lines of commercial products.

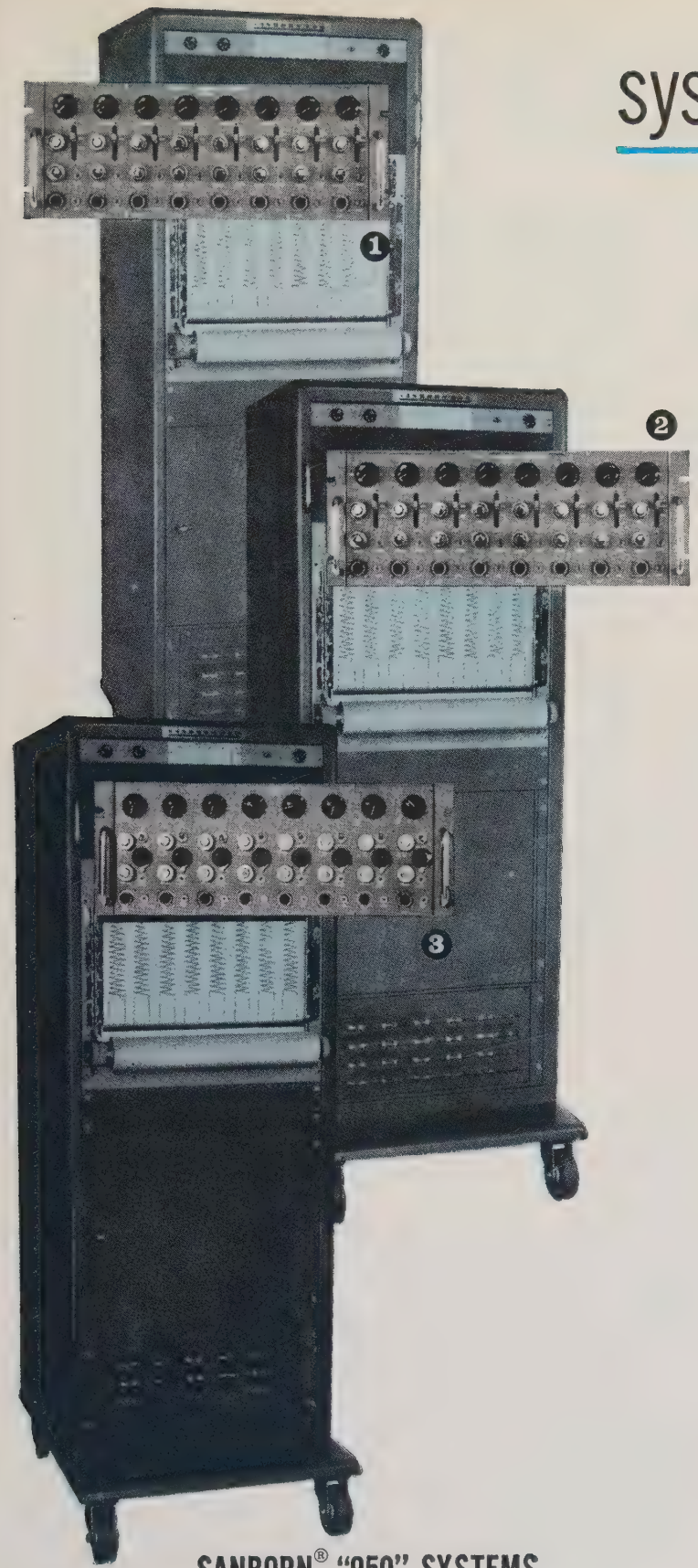
He will continue as Director of the Company's Microwave Tube Division, a post he has held for the last 18 months, and as a member of the Company's Technical Board.

He previously served as Associate Director of the Research Laboratories and as Director of the Physics Laboratory.

He collaborated with Dr. J. R. Pierce in developing the first stable traveling wave tube while with Bell Telephone Laboratories. He later served as Professor of Electrical Engineering at Stanford University, Stanford, Calif., where he invented many forms of traveling wave tube propagating structures. He was Professor of Electrical Engineering at California Institute of Technology, Pasadena, from 1952 to 1960.

Dr. Field received the B.S. degree from Purdue University, Lafayette, Ind., in 1939, and the Ph.D. degree from Stanford University in 1942. He was awarded the Eta Kappa Nu outstanding young engineer award in 1949. He is a member of the American Association of University Pro-

(Continued on page 62A)



match system sensitivity to your recording application

When your recording application calls for 6 or 8 direct writing, general purpose channels with identical sensitivity and input electronics, Sanborn "950" design provides a highly useful, economical answer in the precise sensitivity range you need. Choices include systems with:

1 HIGH GAIN AMPLIFICATION . . . 6 or 8 transistorized channels with floating and guarded inputs, 100,000 ohms resistance on all ranges, 10 to 2000 uv/div sensitivity. System response DC to 100 cps within 3 db at 10 div peak-to-peak amplitude. Common mode performance ±200 volts, max., rejection 140 db min. at DC. High gain stability, max. non-linearity 0.5%, low noise and drift. All channels have range, gain, function, position and galvanometer frequency compensation controls.

2 MEDIUM GAIN AMPLIFICATION . . . 6 or 8 transistorized channels with floating and guarded inputs; ½ meg. resistance on mv ranges, 1 meg. on volt ranges; sensitivity 0.5 to 20 mv/div and volts/10 div. System response DC to 150 cps within 3 db at 10 div peak-to-peak amplitude. Common mode voltage ±500 volts, max., rejection 140 db min. at DC. Same controls as High Gain amplifier.

3 LOW GAIN AMPLIFICATION . . . 6 or 8 channels, balanced to ground inputs 5 meg. each side; tube and transistor circuitry; sensitivity 10 to 500 mv/div and 1 to 10 volts/div. System response same as Medium Gain system. Common mode performance ±2.5 volts and 34 db min. rejection on most sensitive range. Amplifier available with or without Calibrated Zero Suppression.

All "950" systems have 350-style 6- or 8-channel flush-front, heated stylus recorder, using Sanborn rectangular-coordinate Permapaper.® Nine chart speeds, timer/marker stylus, built-in paper take-up. Systems housed in 63⅜" high mobile cabinets; amplifier and recorder occupy only 24½" of panel space. For complete descriptive literature and application assistance, call your Sanborn Industrial Sales-Engineering Representative; offices throughout the U. S., Canada and foreign countries.

SANBORN® "950" SYSTEMS

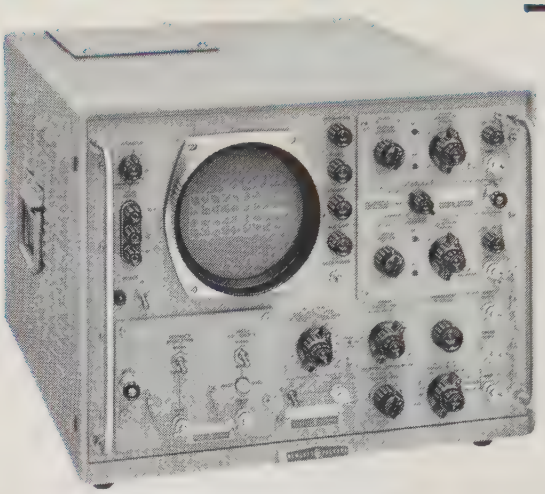
Dynamic demonstrations of 950 Systems and other Sanborn oscillographic recording equipment at WESCON, Aug. 22-25 Booths 2014-2016

INDUSTRIAL DIVISION
SANBORN COMPANY
175 Wyman St., Waltham 54, Mass.

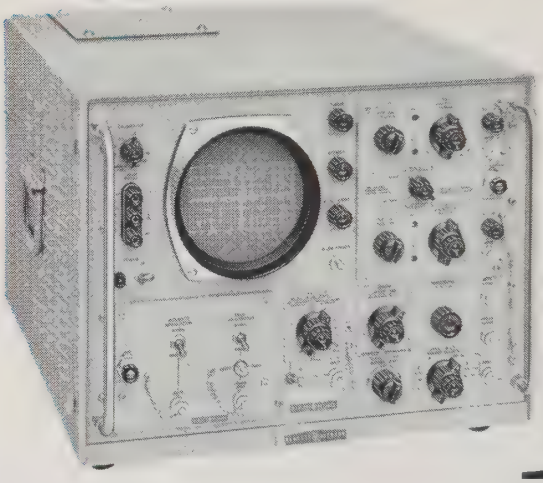
High quality, high reliability plus

PLUG-IN VERSATILITY

hp 170A MILITARIZED SCOPE—TO 30 MC!

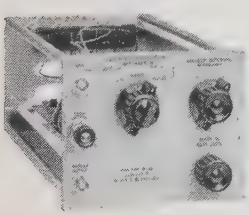


hp 160B MILITARIZED 15 MC SCOPE

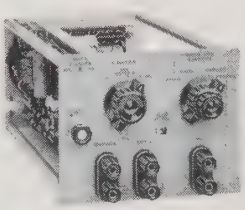


Seven horizontal, vertical plug-ins give

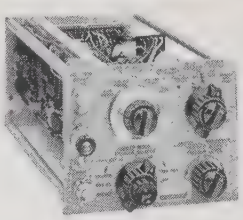
hp 166A Plug-in *(Time-Axis) furnished with the hp 160B and 170A Oscilloscopes (as pictured above), provides standard input connections for Z-axis modulation and single-sweep arming.*



hp 166B Time Mark Generator
(Time-Axis plug-in) makes precise time measurements simple, provides intensity-modulated time markers on the oscilloscope trace of either hp 160B or 170A. Markers, at 0.1, 1 and 10 μsec intervals, speed, simplify photographs, calibration of fast oscilloscope sweeps and operation between calibrated sweep ranges with sweep vernier. Markers may also be used as triggers or for calibration of other devices. Accuracy is ± 0.2%, 10° to 30° C. hp 166B, $130.00.



hp 166C Display Scanner *(Time-Axis plug-in) provides output to duplicate, on an X-Y recorder, any repetitive waveform appearing on CRT trace. Resolution with permanent, large-scale records is higher than either scope CRT or photograph, and you can observe the scope trace while records are made. Unit converts high speed signals to slower signals having the same waveshape; scanning speed is arranged to keep Y output within the bandwidth of conventional recorders. hp 166C, $300.00.*



hp 166D Sweep Delay Generator *(Time-Axis plug-in) delays the main sweep of the 160B and 170A Scopes for detailed examination of a complex signal or pulse train. In addition, it offers a unique mixed sweep feature to show an expanded segment of a delayed waveform while still retaining a presentation of earlier portions of the waveform. Delay time 1 μsec to 10 sec. Delaying sweep, 18 ranges. Delayed length 0 to 10 cm. Delay functions: trigger main sweep, arm main sweep, mixed sweep. hp 166D, $325.00.*

in these hp OSCILLOSCOPES

Vertical or horizontal plug-ins make possible

- Dual trace amplification
- Fast pulse amplification
- High gain amplification
- X-Y records of repetitive waveforms
- New sweep delay convenience
- Time markers for photos, calibration

Both oscilloscopes are highly ruggedized; both have conventional controls for simple, swift operation

Built to exacting military specifications, these hp oscilloscopes offer instantly expandable measurement capability—*when you need it*. It's easy! Just add a moderately priced plug-in unit!

Both hp 160B and 170A employ the same vertical and time-axis plug-ins providing the widest range of application with minimum investment.

New hp 160B and 170A meet MIL specifications for shock, vibration, humidity and temperature. Important features include high stability tube/transistor circuits, regulated dc filament voltages and premium components throughout.

Etched circuits on translucent epoxy glass provide great mechanical stability and simplify circuit tracing. Improved preset triggering insures optimum operation for almost all conditions with just one adjustment—even on signals down to 2 mm deflection. A push-button beam finder automatically locates an off-screen beam or trace, especially important during operation by inexperienced personnel.

SPECIFICATIONS—hp 160B and 170A with hp 166A Plug-in

VERTICAL

Bandwidth:	hp 160B, > 15 MC; hp 170A, > 30 MC
Voltage Calibrator:	18 calibrated ranges ± 3%, 0.2 mv to 100 v peak to peak

HORIZONTAL

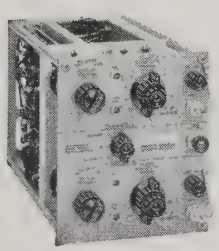
Bandwidth:	DC to 1 MC
Sensitivity:	7 ranges 0.1 v/cm to 10 v/cm. Vernier extends minimum sensitivity to 25 v/cm
Input Impedance:	1 megohm shunted by 30 pf

SWEEP GENERATOR

Internal Sweep:	24 ranges, 0.1 μsec/cm to 5 sec/cm, ± 3%. Vernier extends slowest sweep to 15 sec/cm
Magnification:	7 ranges, X1, X2, X5, X10, X20, X50 and X100. Increases fastest sweep to 0.02 μsec/cm
Triggering:	Internal, power line or vertical input signal (2 mm or more vertical deflection); external (½ v peak to peak or more). Trigger level of external sync signal adjustable –30 to +30 volts
PRICE:	hp 160B, $1,850.00 (cabinet or rack mount) hp 170A, $2,150.00 (cabinet or rack mount)

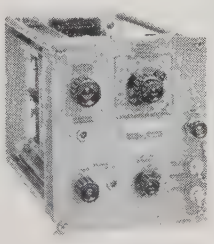
hp 160B, 170A unmatched usefulness

hp 162A Dual Trace Amplifier



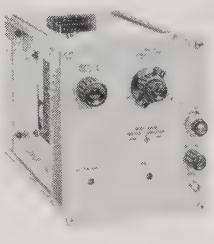
plug-in (vertical) gives maximum sensitivity to 20 mv/cm, permits viewing of two phenomena simultaneously, offers differential input for common mode rejection. Electronic chopping extends simultaneous viewing of 2 signals to lower frequencies without flicker. hp 162A, $350.00.

hp 162D High Gain Vertical Amplifier



increases sensitivity to 5 mv/cm. 12 calibrated ranges, 5 mv/cm to 20 v/cm in 1, 2, 5, 10 sequence, accuracy ± 5%. Continuous vernier extends min. sensitivity to 50 v/cm. Differential input with at least 40 db common mode rejection included for ranges 5 mv/cm through 50 mv/cm. AC or dc coupling of either of two inputs. hp 162D, $225.00.

hp 162F Fast Rise preamplifier



Vertical plug-in allows full utilization of the excellent transient response of the 160B and 170A main vertical amplifiers. Rise time with hp 170A is 12 nsec, dc to 30 MC; with hp 160B is 23 nsec, dc to 15 MC. Sensitivity is 0.05 v/cm to 50 v/cm, covered in 9 ranges; input impedance 1 megohm with 25 pf shunt. hp 162F, $145.

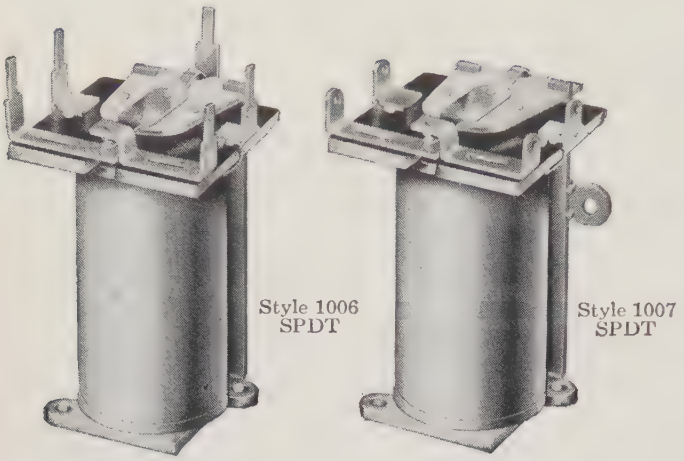
Data subject to change without notice. Prices f.o.b. factory.

HEWLETT-PACKARD COMPANY
7268D Page Mill Road Palo Alto, California, U.S.A.
Cable "HEWPACK" DAvenport 6-7000
Sales representatives in all principal areas

HEWLETT-PACKARD S. A.
Rue du Vieux Billard No. 1 Geneva, Switzerland
Cable "HEWPACKSA" Tel. No. (022) 26. 43. 36

7268

Sensitive Relays at Sensible Prices



Price Electric Series 1000 Relays Now Feature . . .

Sensitive Operation • Solder or Printed Circuit Terminals
Open or Hermetically Sealed Styles • Low Cost

These versatile sensitive relays are designed for applications where available coil power is limited. They retain all the basic features, such as: small size, light weight and low cost, that make the Series 1000 General-Purpose Relays pace setters in their field.

Typical Applications

Remote TV tuning, control circuits for commercial appliances (including plate-circuit applications), auto headlight dimming, etc.

General Characteristics

Standard Operating Current:
1 to 7 milliamps DC at 20 milliwatt sensitivity

Maximum Coil Resistance: 16,000 ohms

Sensitivity:
20 milliwatts at standard contact rating; 75 milliwatts at maximum contact rating. Maximum coil power dissipation 1.5 watts.

Contact Combination: SPDT

Contact Ratings:
Standard 1 amp; optional ratings, with special construction, to 3 amps. Ratings apply to resistive loads to 26.5 VDC or 115 VAC.

Mechanical Life Expectancy:
30,000,000 operations minimum.

Dielectric Strength: 500 VRMS minimum.

For Additional Information, contact:

PRICE ELECTRIC CORPORATION

300 Church Street • Frederick, Maryland
MOnument 3-5141 • TWX: Fred 565-U

IRE People

(Continued from page 58A)

fessors, the American Physical Society, Sigma Xi, Tau Beta Pi and Eta Kappa Nu.

❖

Donald L. Garner (A'60) has been appointed a District Sales Manager for Motorola Semiconductor Products, Inc., it was recently announced.

D. L. GARNER

In his new capacity, he will be responsible for the firm's semiconductor component sales in an assigned area of the midwest. His headquarters will be located in Motorola's Area Office in Chicago, Ill.

Before joining Motorola, he was a Sales Engineer for Hughes Aircraft and Audio Devices, Inc. From 1953 to 1958, he was associated with Motorola, Inc., as Assistant Department Head in the Television Electrical Model Shop.

Mr. Garner is a graduate of the American Television Institute.

❖

The appointment of **George W. Hart** (M'59) as vice president, marketing, of the Electronic Energy Conversion Corp. of Bethpage, L. I., N. Y., was recently announced. Hart comes to Electronic Energy from General Electric Co., where he was manager of marketing administration and personnel development for the light military electronics department. Before this position, he headed up activities at General Electrics' manager development operations at Crotonville, N. Y.

G. W. HART

A graduate of Bucknell University, with a B.S. in electrical engineering, he has been actively engaged in the field of marketing for the past 30 years.

Mr. Hart holds a professional engineering license from the State of New York. He also studied economics at the State Graduate School of New York University. He is a member of the American Institute of Electrical Engineers, American Management Association and the Society for Advancement of Management

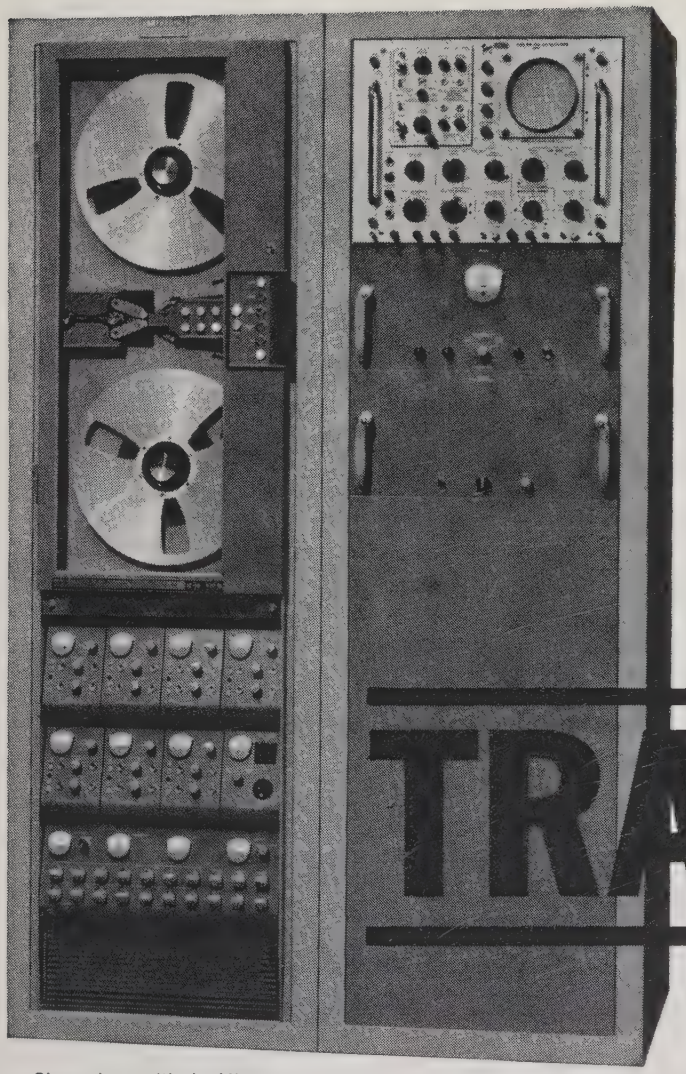
❖

Douglas M. Heller (SM'57) has been appointed assistant general manager at the radio division of The Bendix Corporation, it was announced recently. He also has been named manager of the division's government products group. Prior to his new

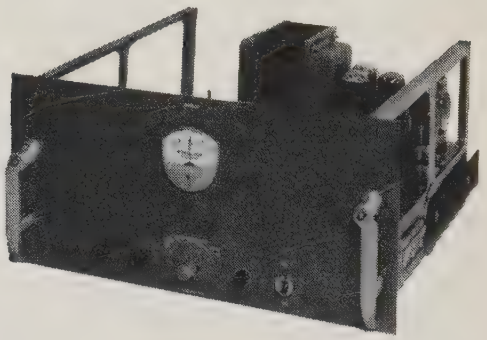
(Continued on page 65A)

Data loss due to drop-outs is eliminated in FM predetection recording-reproducing by Mincom's new, exclusive Tracklok®

This is because, for the first time in the field of instrumentation, the **Tracklok** makes possible redundant FM data recording at the carrier level. In any desired FM or PM-type carrier system, data loss is eliminated by a 99% skew reduction; existing skew of ± 0.3 μs for example, is effectively reduced to ± 0.003 μs, a reduction of 100 to 1.



Shown here with the Mincom Series CM-100 1.5-mc Instrumentation Recorder/Reproducer, a standard auxiliary rack houses (from the top down) an oscilloscope monitor unit, the new **Tracklok**, and a demodulator.



The Tracklok

Completely Compatible: The new **Tracklok** is designed to improve the predetection performance of Mincom's 1-mc Series CM-100 Instrumentation Recorder/Reproducer (which now, on special order, performs to 1.5 mc at 120 ips). **Tracklok** can be incorporated into all existing Series CM-100 systems, since it is compatible with CM-100 or any comparable recorder-reproducer in the standard IF carrier frequencies.

TRACKLOK®

Reliable Simplicity: The same reliability that has been typical of Mincom's instrumentation systems for years has been built into **Tracklok.**

MINCOM DIVISION **3M** **MINNESOTA MINING & MANUFACTURING CO.**

2049 SO. BARRINGTON AVE., LOS ANGELES 25, CALIFORNIA • 529 PENN BLDG., 425 13th ST. N.W., WASHINGTON 4, D.C.

How to turn a new face into an old hand

New employee's learning time cut 75%!
Production increased 50%!
Rejects reduced 55%!

Today at Hughes these levels of improvement have become the norm in manufacturing. New employees who don't know a transistor from a thimble become skilled assemblers in days instead of weeks. They start almost immediately to help build some of the most complex electronic systems ever developed.

More importantly, they can build them right the first time. And they can meet tight schedules while keeping costs down.

The secret is Videosonic* systems, developed and perfected in Hughes plants. Videosonic systems apply proven audio-visual methods in new and highly developed ways. Recorded instructions and slides guide workers step by step. They *see* what they are to do, as they *hear* how to do it. Each work station is integrated into over-all operations—pace is maintained, flow is smoothed.

Videosonic systems also give immediate measurable results in other areas where efficiency is demanded. Clerical workers can quickly learn new methods. Product testing can be speeded. Engineering activities can be made more effective.

Videosonic systems have proven their merit in 5 years of use in Hughes plants in El Segundo and Fullerton, California and Tucson, Arizona. Today, they assure the reliability and speed production of Polaris missiles, electronic armament systems, Falcon missiles, electronic air defense control systems, and advanced computer systems.

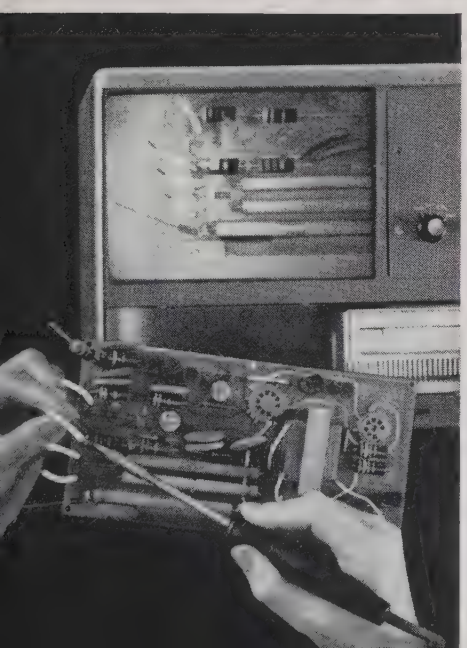
Interested in Videosonic systems for your operation? Hughes has prepared a special booklet detailing the advantages of Videosonic systems. For your copy, please contact: Manager, Videosonic Systems, Hughes Ground Systems Group, Building 381-N, Fullerton, California.

Creating a new world with electronics

HUGHES

HUGHES AIRCRAFT COMPANY

*Videosonic is a trademark of Hughes Aircraft Company



IRE People

(Continued from page 62A)

appointment, he was assistant group executive in charge of the aircraft and automotive sections of the company's Products division, South Bend, Ind.

Mr. Heller, who was born in Dover, England, attended the Royal College of Science in London and was awarded the ARCS and B.S. (physics) degrees. He came to the United States in 1944 for the British government. He joined the radio division in 1947 and later became chief engineer for commercial products engineering. In 1954, he became director of engineering at the Bendix Products missile section (now the Bendix Mishawaka division), Mishawaka, Ind., and later was named assistant general manager of the section. In 1959 he was named assistant group executive at the Products division.

He is a Fellow of the British Institute of Physics, a member of the American Rocket Society, and a member of the Guided Missile Council of the A.I.A.

❖

The promotion of **Dr. Rudolf E. Henning** (A'46–M'52) from Engineering Manager to Chief Engineer of the Sperry Microwave Electronics Company, Clearwater, Fla., has been announced.

R. E. Henning

He directs, coordinates and is responsible for all efforts of the engineering departments at Sperry Microwave. The work involved ranges from basic and applied research through planning for large scale production.

Projects now under way in the engineering departments include basic and/or applied research in the fields of microwave solid state materials and devices, microwave radiometry, advanced measurement techniques, modern antenna systems and special instrumentation studies. Design, development and manufacturing follow-up programs encompass microwave solid-state devices, microwave components, manual and automatic checkout equipment, custom-built support equipment and modules, antennas and Microline test and measuring equipment for laboratory and production line use.

With Sperry since 1947, Dr. Henning has been responsible for the engineering management of the Sperry Rand Clearwater operation since its inception in 1957. He received the D. Eng. Sci. degree from Columbia University, New York, N. Y., in 1954. He has written many technical papers and holds a number of patents. Now Chairman of the Florida West Coast Section of the IRE, he is also a member of numerous other technical associations.

❖

The appointment of **Paul C. Ely, Jr.** (M'56) as Engineering Head of Microline Operations has been announced by the Sperry Microwave Electronics Company, Clearwater, Florida.

P. C. Ely, Jr.

The entire line of Microline laboratory and factory precision measuring instruments now is being handled by a separate, unified organization within the company, which is responsible for every aspect of the Microline operation. Ely heads a team which is responsible for all engineering phases of Microline, from initial concept through setup and tooling for manufacture. Production and marketing groups complete the integrated Microline organization.

Ely, who received his B.S. in Engineering Physics from Lehigh University, has been with Sperry since 1953, specializing in the development, design and marketing of commercial and military test equipment. A member of the PGMTT of IRE, he is also active on the Calibration and Measurements Project Team of the Quality Control Committee of the Aerospace Industries Association.

❖

David R. Hull (A'36–F'47) has been elected an Associate of Boyden Associates, Inc. He was, until recently, Executive Vice President of Hoffman Electronics Corporation in Los Angeles.

D. R. Hull

Mr. Hull has been active in the electronics industry for 35 years as a naval officer and business executive. A graduate of the U. S. Naval Academy, he served for 23 years in naval electronics engineering and administration, retiring with the rank of Captain and as Assistant Chief of the Bureau of Ships for Electronics, the senior navy position in electronic materiel. Prior to World War II, he specialized in UHF radio, underwater sound and radar development. He received the Navy Commendation Ribbon and Citation for his coordination of the Navy's pre-war radar research and development program, and the Legion of Merit for his wartime direction of the design and development of all naval electronic equipment.

Following his retirement from the Navy in 1948, he became Assistant Technical Director of International Telephone and Telegraph Corporation and Vice President and Director of its Federal Telecommunications Laboratory. When ITT acquired Farnsworth, he was elected Executive Vice President and Director of the newly-formed Capehart-Farnsworth Corporation. In 1950 he joined Raytheon Company as Vice President and General

(Continued on page 66A)

CTS

For Small Space Printed Circuit Applications

NEW SUBMINIATURE SINGLE TURN COMMERCIAL TRIMMER POTENTIOMETER

7/16" x 5/16" x 1/2"

- An economically priced unit containing the same composition element, double wiper contactor construction and all other performance-proven quality features of CTS miniature composition variable resistors.
- Requires only 7/16" x 5/16" p. c. board area. Extends only 1/2" above board surface. Plugs directly into board and is self-supporting. Can be used in applications where multiple boards are stacked on 5/8" centers.
- Designed for communications, computers, instrumentation, electro-medical and other small space printed circuit equipment applications.

SPECIFICATIONS for SERIES 220

Resistance Range: 250 ohms thru 2.5 megohms, linear taper. Other tapers available.
Wattage & Temperature Rating: 1/8 watt at 55°C derated linearly to no load at 85°C with control mounted on p. c. board.

Voltage Rating:
Shaft to Terminals: 750 VAC for 1 minute high pot test, 500 VDC operating maximum.
Across End Terminals: 350 VDC. Load not to exceed wattage rating.

Angle of Rotation: 300° ±5°.

Request Data Sheet 184 containing complete technical description.

For your military, industrial and commercial applications, CTS manufactures the world's greatest variety of variable resistors . . . both composition and wire wound. Draw upon the expert knowledge and willing help offered by CTS variable resistor specialists.

FOUNDED 1896

CTS CORPORATION
ELKHART, INDIANA

Factories in Elkhart & Berne, Indiana; South Pasadena, California; Asheville, No. Carolina and Streetsville, Ontario, Canada. Sales offices and Representatives conveniently located throughout the world to help solve your variable resistor problems.

IRE People

(Continued from page 65A)

Manager of Equipment Operations, retiring on June 1, 1960. On September 1, 1960 he joined Hoffman Electronics Corporation as Executive Vice President.

Mr. Hull served as President of the Electronic Industries Association from 1958 to 1960, and was awarded the Association's 1960 Medal of Honor for outstanding contributions to the industry. In addition to the Bachelor of Science degree, Mr. Hull holds an M.S. degree from Harvard University. He has been awarded fellowships in the Institute of Radio Engineers and the Acoustical Society of America, and holds membership in several other professional societies. He is a Director of Mycalex Corporation of America, Armed Forces Communications and Electronics Association, and the American Standards Association.

❖

The new corporate position of Product Exploitation Director has been established at Hughes Aircraft Company, Culver City, Calif. **Lance P. Johnson** (SM'58) has been named to the post. He will work with all divisions of the company to evaluate products of Hughes' total technical effort and develop additional applications for them where possible.

L. P. Johnson

As Product Exploitation Director he will report to L. A. Hyland, Vice President and General Manager. In addition, he will retain his position as Manager of the Industrial Systems Division and also manage a newly-created Electronics Products Division of the Components Group. He will make his headquarters at the Hughes Newport Beach, Calif., plant.

He joined Hughes in 1951 from Bell Telephone Laboratories, where he was a number of the technical staff. At Hughes he has been Associate Director and Director of Operations for the Airborne Systems Laboratories, and Assistant Division Manager and Director of Operations for the company's El Segundo, Calif., manufacturing division.

Mr. Johnson studied electrical engineering at Rensselaer Polytechnic Institute, Troy, N. Y., and attended the executive program of UCLA graduate school of business administration. He is a member of the Scientific Research Society of America.

❖

The appointment of **William D. Hogan** (S'50–A'51–M'55) to the newly-created position of Manager of Engineering Services for the Semiconductor Division of Sylvania Electric Products Inc., has been announced.

He has been Manager of Field Engi-

(Continued on page 68A)

HIGH PURITY ELEMENTS

SOUTH PLAINFIELD, N. J., Oct. 1 — ASARCO CENTRAL RESEARCH LABORATORIES, USING SPECIAL REFINING AND ANALYTICAL TECHNIQUES, HAVE SUCCEEDED IN PURIFYING SELENIUM SO THAT NO NICKEL OR IRON CAN BE DETECTED AT 5 PARTS IN 100,000,000. LEVELS OF OTHER IMPURITIES IN ASARCO HIGH PURITY ELEMENTS CAN AT TIMES BE MORE SPECIFICALLY LIMITED TO MEET CUSTOMERS' SPECIAL REQUIREMENTS.

ASARCO HIGH PURITY ELEMENTS ANALYSES as of OCT. 1, 1961

Impurities sometimes found at maximum levels in parts per million
(< denotes less than ppm indicated)

	Element and Grade		Bi	Cu	Fe	As	Pb	Ag	Tl	Sn	Te	Au	Na	Cl	Cd
	ANTIMONY	A-60	<1	<1	<1	<2	<1								
✓	ARSENIC	A-60		<1											
	BISMUTH	A-58		2	1		1	2						<1	
	CADMIUM	A-60		<1	<1		1								
	COPPER	A-58	No Impurities Detectable by Spectrographic Analysis												
	GOLD	A-59		<1	<1		<1	1							
	INDIUM	A-58		<1			1		<1	1					
	LEAD	A-59	<1	<1	<1										
	SELENIUM	A-58		<1		<1					1			<1	
	SILVER	A-59		<1	<1		<1								
	SULFUR	A-58											1	1	
	TELLURIUM	A-58		<1	<1										
✓	THALLIUM	A-60		1	1		<1	<1					2	<1	
	ZINC	A-59		<1	<1		<1								<1

✓ *Denotes element improved since last published analyses.*

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(Continued from page 66A)

neering since June, 1960, and will now coordinate all engineering activities associated with marketing. He joined Sylvania in 1950 as a Junior Engineer at the division's Engineering Test Laboratory in Boston. Prior to joining Sylvania, he was with Raytheon Manufacturing Company at Waltham, Mass.

A native of Cambridge, Mass., he graduated summa cum laude from the University of Massachusetts, Amherst, Mass., where he received the Bachelor of Science degree in electrical engineering. He has also followed courses in industrial electronics at Northeastern University, Boston, Mass.

Mr. Hogan is a member of the IRE Professional Group on Electron Devices, a member of JEDEC JS-2, Committee on Signal Diodes, and a Technical Associate to the Aerospace Industries Association, Semiconductor Device Panel.

❖

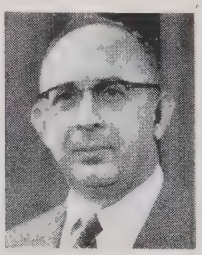
Walter F. Joyce (SM'56) has been appointed Senior Vice President of Texas Instruments, Inc., Dallas, Tex., with responsibility for the Apparatus Division and additional corporate responsibilities to be assigned.

He joined the company in 1949 as Manager of Manufacturing, was elected a Vice President in 1951, and assigned responsibility for the Apparatus Division upon its creation in 1953. He received the Bachelor's and Master's degree in chemistry and physics at Marquette University, Milwaukee, Wis., in 1932 and 1943, respectively. During World War II he was in the Navy's Bureau of Aeronautics, and was discharged with the rank of Lieutenant Commander.

Mr. Joyce is a Senior Member of the American Chemical Society and a Trustee of the National Security Industrial Association. His other memberships include the American Ordnance Association, the Armed Forces Communications and Electronics Association, and the Electronics Industries Association.

❖

The appointment of **Louis Kahn** (A'31–SM'51) as Vice President in Charge of Engineering for Astron Corporation, East Newark, N. J., and its subsidiaries has been announced.



L. Kahn

Well known in the electronics industry for his major contributions in the electronic components field, he is recognized as one of the foremost authorities on capacitors and associated components.

Formerly associated in management and engineering capacities with Aerovox

Corporation for over twenty years, he pioneered the development of miniature and subminiature components for many advanced electronic applications. Prior to his position as Director of Engineering at Cinema Engineering, Division of Aerovox Corporation, he was Technical Assistant to the President of Aerovox Corporation, Director of Engineering and Research, and Assistant Vice President of Technical Sales. In addition to his many accomplishments at Aerovox Corporation, he has actively participated in key industry and government activities as the Chairman of the Subpanel on Capacitors of the Department of Defense, Chairman of Technical Committee on Fixed Capacitors of the Electronics Industries Association, Vice Chairman of the Professional Group on Component Parts of the IRE, Chief Technical Advisor for the USA Technical Committee on Capacitors of the Industrial Electrotechnical Components Committee, and many other important electronic industry groups.

As an authority on dielectric materials for electronic components, he has lectured at Massachusetts Institute of Technology, Cambridge, and has written over fifty technical papers on capacitors and related subjects.

Mr. Kahn is a graduate of Rutgers University, New Brunswick, N. J., with the degree of Bachelor of Science and Master of Science in electrical engineering and taught electrical engineering at Rutgers for ten years.

In his new capacity as Vice President in Charge of Engineering at Astron Corporation, he will direct the company's research and development programs, administrate all engineering activities, and participate on the management team of the corporation and its subsidiaries.

❖

The appointment of **Robert L. Kent** (M'59) to the new position of Manager of the Electronic Systems Engineering Department for Itek Electro-Products Company, Cambridge, Mass., has been announced.

Itek Electro-Products, a division of Itek Corporation, is a producer of crystal filters and related devices.

In his new position, Mr. Kent will be responsible for the engineering phases of larger military and commercial systems based upon the growing application of crystal devices. Itek spectrum analyzers, extensively used in space surveillance systems, are such devices.

He joined Itek in January, 1961, after ten years with the Research Laboratory of Electronics at M.I.T., Cambridge, where he was a staff member and leader of a group investigating techniques for the measurement of the gravitational red shift postulated by Einstein's General Theory of Relativity. He is co-author of several papers on this subject.

He has the B.S. degree in electrical engineering from the University of Pennsylvania, Philadelphia, and the S.M. degree from M.I.T. He is a member of Eta Kappa Nu, Sigma Tau, Tau Beta Pi, and is an Associate of Sigma Xi.

❖

(Continued on page 70A)

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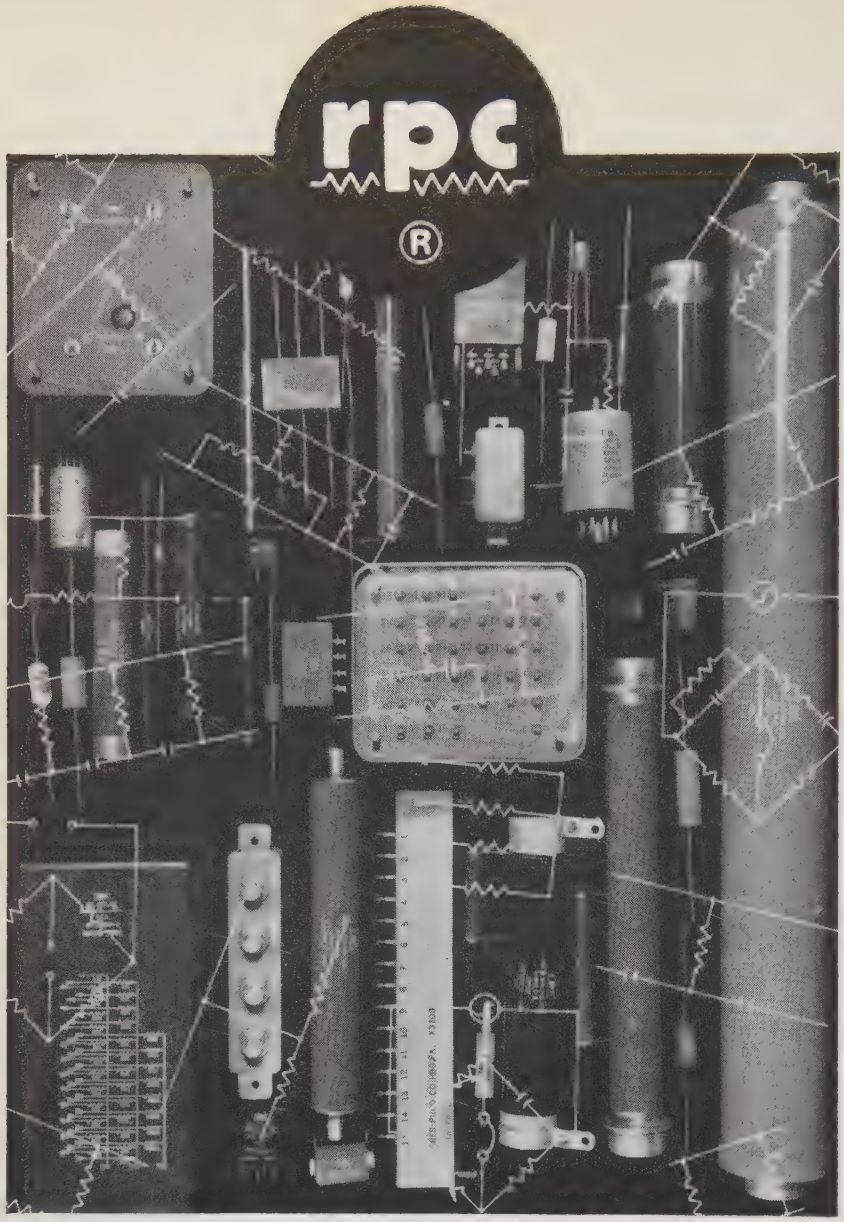

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(Continued from page 69A)

The appointment of **John C. Keyes** (M'49–SM'56) to the position of Vice President of Engineering of the newly-formed Jensen Engineering, Inc., Palo Alto, Calif., has been announced.

J. C. KEYES

He was formerly with the Philco Western Development Laboratories in Palo Alto, where he was Director of the System Engineering Laboratory.

He received the M.S. degree in engineering from the University of California at Los Angeles. He spent 12 years at the U. S. Naval Ordnance Test Station, China Lake, Calif., where he was responsible for the development of airborne radar and fire control systems. In 1958 he joined the Philco Corporation, where he held a position of major responsibility in that firm's activities in communication and control for Air Force space programs.

❖

The appointment of **Robert G. Kinsman** (S'53–M'56) to the new position of Manager, Crystal Filter and Components Engineering Department, for Itek Electro-Products Company was announced recently. Mr. Kinsman was Director of Engineering for Electronic Laboratories Corporation of Torrance, Calif., when that crystal filter-producing firm was acquired recently by Itek Corporation. He had previously been associated with the Crosley Division of Avco, Hughes Aircraft, and the U. S. Air Force, where he was an instructor at the U. S. Air Force Radar Technical School.

He has a degree in electrical engineering from the University of Cincinnati and a master's degree from the University of Southern California. He is a member of Tau Beta Pi, Eta Kappa Nu, and Phi Eta Sigma.

❖

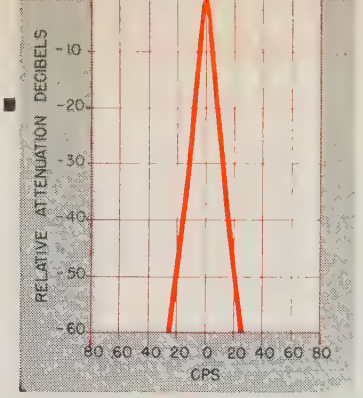
Walter A. Kirsch (SM'57) has been named Assistant to the President of Elm Instrument Corp., Hempstead, L. I., New York electronics firm. The company designs and produces accelerometers, air, mechanical, and magnetic clutches, variable adapters, and other devices for industrial and military uses. The position is a new one.

W. A. KIRSCH

Before joining Elm Instrument, he had been assistant to the president of ID

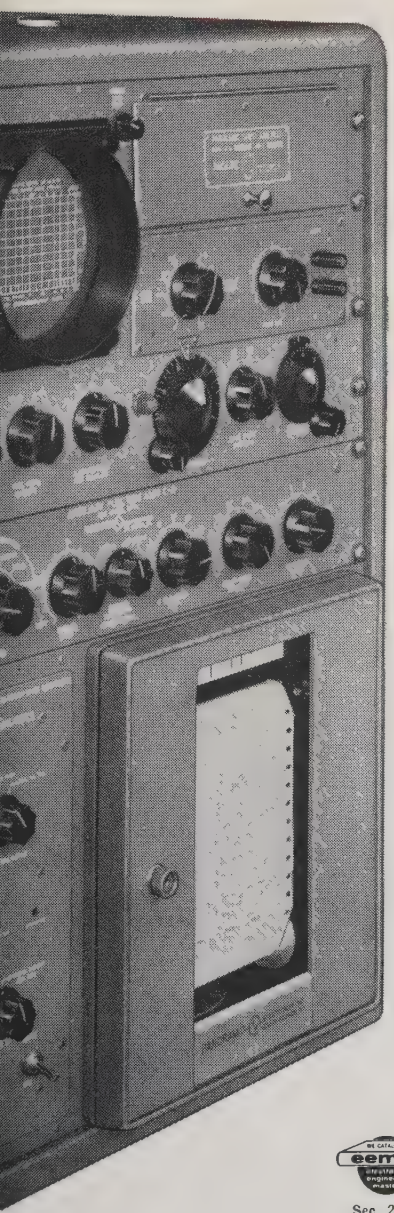
(Continued on page 72A)

now see this...

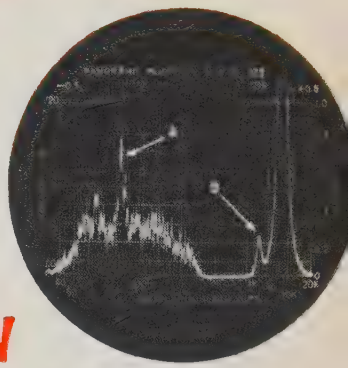


Bandpass response for 2 cps I-F bandwidth

2 cps resolution from 5 cps through 20 kc



with Panoramic's LP-1a SONIC SPECTRUM ANALYZER IN COMBINATION WITH THE NEW C-2 AUXILIARY FUNCTION UNIT.



Narrow I-F bandwidths with steep skirts enable:

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b) amplitude measurement of weak signal closely adjacent to very strong signal.

Panoramic's new low cost system, consisting of the famed Model LP-1a Sonic Analyzer, new C-2 Auxiliary Function Unit and RC-3b Recorder, is designed to solve an extremely wide range of noise, vibration and other waveform analysis problems.

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Direct frequency and amplitude readouts of discrete and random waveform contents are fast and accurate . . . simplifying use at production test stations or in the laboratory.

Indeed, whether your need is to analyze random or line spectra either with components deep in noise or widely apart in level but close in frequency, this new Panoramic system is your assurance of faster, easier solutions.

Among the system's outstanding capabilities:

- **Overall Frequency Range:** 5 cps to 22,500 cps.
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- **Selectivity Characteristics:** Automatic optimum (second scan) plus adjustable I-F bandwidths from 2 cps to 200 cps with extremely steep skirts—*see graph above* —closely approaches ideal rectangular filter desired in random signal analysis.
- **Read out:** 5" CRT and 12" x 4½" inked chart (with RC-3b
- **Amplitude scales:** Calibrated linear and 40 db log.
- **Residual distortion:** Dynamic range better than 60 db
- **Analyzer stability:** less than 5 cps/HOUR drift (after warm-up) assures meaningful analysis with 2 cps resolution
- **Sensitivity compensation, random and discrete:** Sensitivity remains constant for all I-F bandwidth settings for either random or discrete input signals.
- **Adjustable smoothing time constant:** Smooths random fluctuations, clearly displays average noise level at each frequency for faster analysis.
- **Manual scan control:** CRT beam and analyzer tuning positioned anywhere within sweep width range for extended time vs. amplitude studies. Amplitude vs. time analysis also provided with RC-3b Recorder.

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(Continued from page 70A)

Precision Components Corp., Jamaica, N. Y. and Defense Products Manager of Telechrome Manufacturing Co., Amityville. Earlier he had been with Servo Corporation of America and with Fairchild Camera and Instrument Company. He is a director of the New York Chapter of the Armed Forces Communications and Electronics Association.

❖

C. W. Creaser, Jr., President of Antenna Systems, Inc., Hingham, Mass., announced that **Kendrick H. Lippitt** (M'46 SM'55) has been appointed Vice President, and will be in charge of the new ASI Pacific Division.

K. H. Lippitt

He has been associated with the antenna industry for many years, and comes to ASI from Rohr Aircraft Corporation, where he was Manager of the Antenna Systems Division. He has also been connected with Hughes Aircraft Company as Manager of the Large Antenna Department; D. S. Kennedy and Company as Sales Manager of Commercial Antenna Systems; and Technical Appliance Corporation as Vice President of Engineering.

Mr. Lippitt received the B.S. degree from Massachusetts Institute of Technology, Cambridge, has taken graduate courses at Cornell University, Ithaca, N. Y., and attended the Naval radar schools

❖

Ladis D. Kovach (A'53–M'58), professor of mathematics and acting head of the Department of Mathematics and Physics at Pepperdine College has received a part-time appointment as visiting professor at the University of California at Los Angeles. He will work with the Curriculum Study Committee in the Department of Engineering. Dr. Kovach was formerly with Douglas Aircraft Company.

❖

Robert C. Lockwood (SM'57), former Vice President of Instruments for Industry, Inc., Hicksville, N. Y., an independent developer of electronic countermeasure systems, precision potentiometers and related components, has been named Executive Vice President, it was recently announced. He is also a Director of the company. He joined IFI in 1953 as Vice President in charge of engineering, and has been in charge of the over-all direction and planning of the company's development programs. Before joining IFI, he was an engineer with the General Electric Company and Airborne Instruments Laboratory, Inc.

R. C. Lockwood

Mr. Lockwood holds the Bachelor of Science degree in electrical engineering from Clarkson College of Technology, Potsdam, N. Y.

❖

Dr. Frederick B. Llewellyn (A'23–F'28), assistant to the president at Bell Telephone Laboratories, will join The University of Michigan Institute of Science and Technology (IST) as research physicist, effective September 1. He will be assigned to the office of the director to assist in current research direction and in long-range planning of IST research. He also will do research in his special areas of interest.

In announcing the appointment, Dr. J. A. Boyd, director of the Institute, said "Dr. Llewellyn has had a distinguished 35-year career with the Bell Telephone Laboratories and is well known in the fields of communications and electron physics. Many of his contributions to science are documented in the 19 articles he has published in professional journals. In January of this year, he lectured in Ann Arbor as part of IST's World Space Communications Series.

"Dr. Llewellyn has also served the nation as scientific consultant to the secre-

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tary of war and secretary of defense, and as executive secretary to the President's Scientific Advisory Committee."

Dr. Llewellyn received a master's degree at Stevens Institute of Technology and a doctorate at Columbia University. He is a former president of the Institute of Radio Engineers and a fellow of the American Physical Society. In 1935 he was given the Morris Liebmann Award of the Institute of Radio Engineers.

❖

Professor William A. Lynch (A'44–SM'52) has been named head of the Department of Mechanical Engineering at the Polytechnic Institute of Brooklyn. He succeeds Professor Charles T. Oergel who has resigned the headship to devote his full time to teaching and research.

An alumnus of the Polytechnic, he received his first degree in mechanical engineering, and worked in industry until the beginning of World War II, serving from 1929 to 1942 as president of A. K. Lynch Inc., Brooklyn. He joined the faculty of the Polytechnic in 1942. He later received his master's and doctor's degrees in electrical engineering from the Polytechnic.

During the war he served as a staff member and technical manager of Polytechnic Research and Development Corporation, producers of microwave attenuators for the government. In the years following the war he served successively as administrative officer of the communications division of the electrical engineering department, co-chairman of the undergraduate division and finally as administrative officer in charge of the department.

Professor Lynch has written three books in collaboration with Dr. John G. Truxal, head of the Electrical Engineering Department. They are "Introductory System Analysis," "Principles of Electronic Instrumentation," and "Signals and Systems in Electrical Engineering."

He is past president of the IRE circuits committee, and is also a member of the American Society for Engineering Education, Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

❖

Louis Martin (M'44–SM'47) has been appointed Director of Marketing at Eitel-McCullough, Inc. it was announced recently. He will direct all domestic and foreign marketing activities for Eimac electron-power tubes and associated accessories.

L. Martin

He has been active in the electronics industry for more than 26 years. Prior to his present appointment, he was Manager of Marketing Operations at Eimac and was responsible for Commercial, Government and Export Marketing, Advertising and Sales Promotion, and Market Research. Before joining Eimac, he was General Marketing Manager of the Westinghouse Electron Tube Division at

(Continued on page 74A)

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20D632-4C	8	26	36 CT	.35	6200	3.1	99,800
20D627-2C	8	115	115/57	.33	6200	3.5	98,000
20D603-2C	11	115	115/57	.60	6200	3.5	43,300
20D603-4C	11	115	36 CT	.60	6200	3.5	43,300
20D590-2A	15	115	115/57	1.3	4800	6.2	27,800
20D612-2C	18	115	115/57	2.3	4800	9.1	31,000

MOTOR TACHS	SIZE	Ø1	Ø2	TACH. V	STALL TORQUE	N. L. SPD	V. 1000	TOTAL NULL	THEOR. ACCEL.
20D628-2C	8	115	115/57	26	.33	6200	.20	.019	75,800
20D631-4C	8	26	36 CT	26	.35	6200	.20	.019	80,500
20D634-2C	8	26	26	26	.22	6200	.20	.019	75,000
20D634-4C	8	26	36 CT	26	.22	6200	.20	.019	75,000
20D604-2F	11	115	115/57	115	.60	6200	.500	.019	32,600
20D604-4F	11	115	36 CT	115	.60	6200	.500	.019	32,600
20D593-2A	15	115	115/57	115	1.3	4800	3.1	.019	17,500
20D614-2C	18	115	115/57	115	2.3	4800	3.1	.019	25,900

INERTIA DAMPED	SIZE	Ø1	Ø2	STALL TORQUE	STALL POWER	N. L. SPD	DAMPING
20D618-2B	8	115	40/20	.30	3.5	6200	40 DYNE CM
20D605-2D	11	115	115/57	.60	3.5	6200	100 DYNE CM
20D613-2C	18	115	115/57	2.3	9.1	4800	750 DYNE CM

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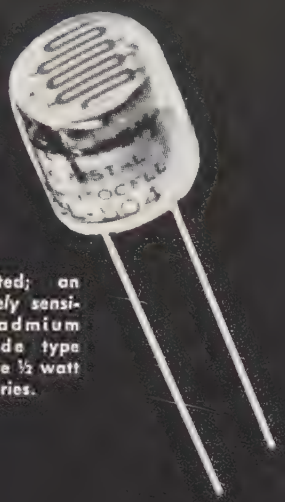

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(Continued from page 73A)

Elmira, N. Y. He served previously for 15 years as Engineer, Application Engineering Manager and Equipment Field Force Manager with the Radio Corporation of America.

He received his bachelor's degree in Electrical Engineering from Cooper Union which was followed by graduate work at Columbia and the University of Pennsylvania. He has served on several committees of the Electronics Industry Association.

❖

The appointment of **Fred L. Martinson** (SM'55), vice president of engineering, as vice president of the newly formed engineering and manufacturing division of Electronic Associates, Inc., has been announced.

F. L. Martinson

The combining of engineering and manufacturing operations under one head was made to speed the transition from the development and engineering to the manufacturing phase of its products. EAI is the country's largest developer and manufacturer of general purpose analog computers, related data processing equipment and instruments.

❖

Jerome L. Nishball (A'50–M'55), formerly Head of Systems Engineering for the Norden Data Systems Division of United Aircraft, has joined MS&A, Inc., as a senior consultant.

With twelve years of experience in the design of electronic computers, Mr. Nishball is one of the pioneers in the field. One of his early projects was the design of input-output equipment for the RAYDAC, one of the first large-scale missile tracking computers.

❖

Eugene B. Novikoff (S'43–A'45–M'51), engineering vice president of Instruments For Industry, Inc., Hicksville, N. Y., has been promoted to the newly established position of vice president of operations. He will be responsible for the over-all management of both the engineering and production departments in IFI's expanded operations.

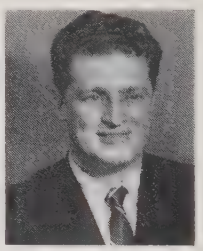
E. B. Novikoff

Mr. Novikoff was one of the first members of IFI's engineering staff, joining the company in April, 1952, as project engineer. He later served as chief engineer and director of engineering and was named a vice president in April, 1959. Earlier, he was a development engineer at Servo Corporation of America and a project engineer at Airborne Instruments Laboratory. He has also been a laboratory instructor in the advanced radio laboratory at the City College of New York, and instructor in the Army Education Program at Ft. Monmouth, and during his military service, was assigned to the Signal Corps Engineering Laboratories at Ft. Monmouth.

A graduate of CCNY, where he received a degree in electrical engineering, he holds a master's degree from the Polytechnic Institute of Brooklyn.

❖

Thomas H. O'Brien (A'50–M'55) has been promoted to the post of vice president —operations of PRD Electronics, Inc., subsidiary of Harris-Intertype Corporation. He will have over-all responsibility for the manufacturing, engineering, research, and weapon systems divisions at PRD.



T. H. O'Brien

He received a B.A. degree in Physics from New York University in 1941, and took additional engineering courses at New York University and

(Continued on page 76A)

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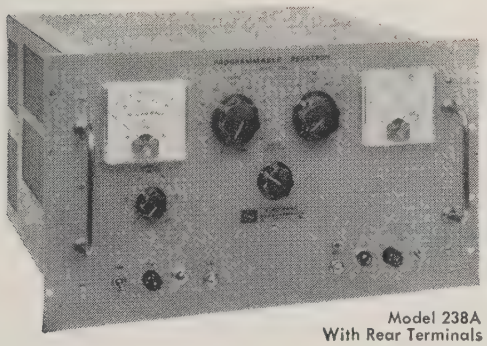
FROM ELECTRONIC MEASUREMENTS

0–1500 V DC POWER SUPPLIES



Model 238A
With Rear Terminals

CONSTANT-VOLTAGE

In addition to covering a complete voltage range—0 to 1500 V—this programmable, regulated power supply can deliver rated current of 100 ma at any output voltage . . . even at less than a volt.

But an even more useful feature is its continuously variable current-limiting control. Here's a feature that's finding more application with every passing day.

When it comes to regulation, the Model 238A is truly flexible. Standard regulation is 0.1% line or load. However, 0.01% regulation is available by means of an optional plug-in chopper unit.

The chart below lists just a few of the many Electronic Measurements off-the-shelf, constant-voltage power supplies. For complete information request Catalog BR1961.

MODEL	RANGE VOLTS DC	RANGE CURRENT
238A	0-1500 V DC	0-100 MA
239A	0-600 V DC	0-1 AMP
235A	0-600 V DC	0-500 MA
234A	0-300 V DC	0-500 MA
230A	0-200 V DC	0-1 AMP
228A	0-150 V DC	0-1 AMP
229A	0-150 V DC	0-300 MA
218A	0-100 V DC	0-3 AMP
221A	0-100 V DC	0-500 MA
225A	0-75 V DC	0-2 AMP
215A	0-50 V DC	0-3 AMP

Model C638A
With Rear Terminals

CONSTANT-CURRENT

Here's the newest in Electronic Measurements' line of Constant Current Power Supplies . . . a full 1500 V compliance at constant current outputs from less than a ma to 100 ma. Like all Electronic Measurements Power Supplies, the Model C638A has a lot of built-in extras too. For example, there's a modulation input . . . there are also provisions for remote programming . . . and an adjustable voltage compliance control that in effect provides for voltage-limiting.

Specifications include a high output impedance of 10^4 megohms at 0.5 μa to 0.5 megohm at 100 ma. Above 2.2 μa, regulation is better than 0.15%, line or load. Ripple is less than 0.01% + 1 μa.

The following listing shows some of the constant current models available. For complete information, request Specification Sheet 3072B.

MODEL	RANGE MIN.	RANGE MAX.	VOLTAGE COMPLIANCE (MINIMUM)
C614A	10 μa	1 AMP	100 V
C613A	10 μa	1 AMP	50 V
C621A	5 μa	500 MA	100 V
C620A	5 μa	500 MA	50 V
C633A	2.2 μa	300 MA	300 V
C629A	2.2 μa	300 MA	50 V
C632A	2.2 μa	220 MA	300 V
C624A	2.2 μa	220 MA	100 V
C638A	1 μa	100 MA	1500 V
C631A	1 μa	100 MA	300 V
C612A	1 μa	100 MA	100 V

ELECTRONIC
MEASUREMENTS
COMPANY, INCORPORATED
EATONTOWN • NEW JERSEY

IRE People

(Continued from page 74A)

Polytechnic Institute of Brooklyn. He is currently attending New York University Graduate School of Business Administration, working toward a degree in management.

Prior to joining PRD in 1959, he was employed by the Sperry Gyroscope Company for several years as an Engineering Department Head in charge of Field Test Equipment and Instrumentation. In this capacity, he was in charge of support equipment for various types of systems, for both Sperry and outside companies, ranging through Fire Control, Navigation, IFF, ECM, Guidance, and Search types.

During World War II, O'Brien was assigned to the Navy as a Field Engineer, instructing on and maintaining carrier aircraft and patrol bomber airborne search and interceptor radar systems. Subsequently he was assigned to the M.I.T. Radiation Laboratory, where he worked with Group 55 on radar test equipment. After World War II, he returned to Sperry, where he headed the Field Test Equipment Group.

During his 20 years of experience in the support equipment field, he has been responsible for the development of many new test equipment concepts and various types of support equipment from portable (TS-147, SPM-6, UPM-32, etc.) to automatic or semi-automatic test systems, such as those used in the B-58 Hustler program.

O'Brien has served on the Executive Committee of the Long Island Section of IRE and on other Institute Committees. He is also a member of the SAE Committee on Military Ground Support Equipment.

❖

The appointment of **Neal Pike** (S'51–A'54–M'60) as Laboratory Director for Prodelin, Inc., was announced recently. In the new position, he will assist Mr. R. F. Lewis, Vice President in charge of Research and Development, in directing new laboratory facilities now under construction at Millstone, New Jersey.

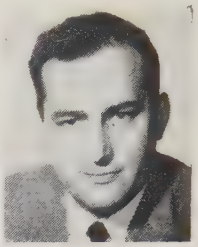
N. Pike

Having served as a member of the Bell Telephone Laboratories and M.I.T. Lincoln Laboratory Technical Staffs, he joins Prodelin, Inc., with a broad background of engineering and production experience with leading companies.

Mr. Pike has a degree in Electrical Engineering from Rutgers University, an M.S. from Cornell University, and is studying for his Ph.D. at Princeton University College of Engineering. He is a member of Tau Beta Pi, Eta Kappa Nu, Pi Mu Epsilon Sigma Xi and The Rutgers Gold Society, and he has received U. S. Navy citations for commendable performance in the submarine service.

❖

Polarad Electronics Corporation announces the appointment of **Herbert W. Pollack** (S'48–A'50–M'55–SM'58) to Engineering Manager.



H. W. Pollack

Joining Polarad in 1955 as Head of the Computer Group, Mr. Pollack moves to his new position from the post of Head of Polarad's General Engineering Section of the Defense Products Division. In that capacity, his responsibilities encompassed general supervision of circuit design, computer techniques, video and RF component development and lumped constant circuitry design.

As Engineering Manager, he will have cognizance over Polarad's military and commercial development projects. His department will also include basic research of microwave components and techniques.

Previous to his Polarad association, he was a senior engineer for CBS, a research associate and staff member at New York University and a design engineer for the A. B. DuMont Laboratories. He holds a B.E.E. degree from C.C.N.Y. and an M.E.E. degree from N.Y.U. Mr. Pollack is presently Chairman of the New York Section of the IRE. He is on the graduate school staff of the Department of Electrical Engineering at Polytechnic Institute of Brooklyn, and is a member of Tau Beta Pi and Eta Kappa Nu, Engineering Honor Societies.

❖

Alex Robbins (M'52) has been appointed Executive Vice President and Director of Research and Development for Standard Systems Corporation, which is engaged in the development, manufacture and marketing of electronic intercommunication systems, industrial control systems and signaling and security systems.

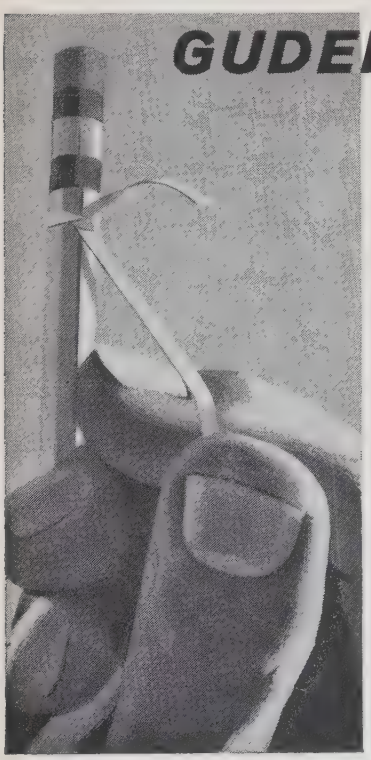
A. Robbins

Prior to this association, Mr. Robbins was employed for fifteen years by Executone, Inc., as Chief Engineer. During World War II, he was with the Bendix Aviation Corp. in the development of U. S. Navy Sonar equipment. He holds a degree in electrical engineering from the Institute of Technology, Vienna, Austria.

❖

Jay R. Reese (S'49–A'50–M'55) has been appointed Vice President in charge of the newly-created Components Division of Texas Instruments, Inc., Dallas, Tex.

He came to the company in 1955 as a product marketing manager in the Semi-



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the lacing tape with a NON-SKID tread

You can't see it, but it's there! Gudelace is built to grip—Gudebrod fills flat braided nylon with just the right amount of wax to produce a non-skid surface. Gudelace construction means no slips—so no tight pulls to cause strangulation and cold flow.

But Gudelace is soft and flat—stress is distributed evenly over the full width of the tape. No worry about cut thru or harshness to injure insulation . . . or fingers.

Specify Gudelace for *real* economy—faster lacing with fewer rejects.

Write for free Data Book. It shows how Gudelace and other Gudebrod lacing materials fit your requirements.

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New York 1, New York

EXECUTIVE OFFICES
12 South 12th Street
Philadelphia 7, Pa.

conductor-Components Division. He subsequently was District and National Field Sales Manager for the Division, and in 1960 was named Components Group Manager. He recently was named General Manager of the newly-created Components Division.

Mr. Reese received the Bachelor's degree in electrical engineering at Princeton University, Princeton, N. J., in 1950. During World War II he served with the Army Signal Corps in the Far East, specializing in radio telephone installation and repair.

❖

Dr. Harry G. Romig (SM'51), previously senior scientist and staff member of Operations Research, Inc., has been named corporate director of quality engineering of Leach Corporation, with primary emphasis on Relay Division activities.

H. G. Romig

Dr. Romig is an international authority on sampling, quality control, reliability, operations research, industrial engineering and management science. He started his research and development work in these fields at Bell Telephone Laboratories, with which he was associated for over 24 years.

He also worked with Donald A. Quarles, former Assistant Secretary of Defense, at BTL on Quality Assurance and many varied research projects. He was awarded the 1953 ASQC Shewhart Medal for his contributions to Quality Control, and in 1949 was classed by *Fortune* magazine as one of four pioneers in the quality control field.

Dr. Romig was graduated from Pacific University in 1921 with an A.B. degree with honors in mathematics. He received an M.A. in mathematics from the University of California, Berkeley, in 1923, and a Ph.D. in industrial engineering from Columbia University in 1939. He has served as a faculty member at the University of California (Berkeley and also at Los Angeles), San Jose State, Brooklyn College, Columbia University and the University of Southern California.

Dr. Romig has also lectured on guided missile engineering and production at a special USC-NATO course in Paris in 1958. Since 1948 he has served as a member of the Advisory Committee for Engineering at Columbia University and is now Adjunct Professor of Industrial Engineering at the University of Southern California.

He is the author of two books and scores of scientific papers but is best known for the classic Sampling Inspection Tables written jointly with Harold F. Dodge.

As Senior Scientist at Operations Research, Inc., Dr. Romig acted with the ORI Project Leader and the Special Projects group in the Navy on means and methods for monitoring and evaluating the reliability of the Polaris project, including preparation of a Problem Manual for PERT-PEP Systems.

(Continued on page 78A)

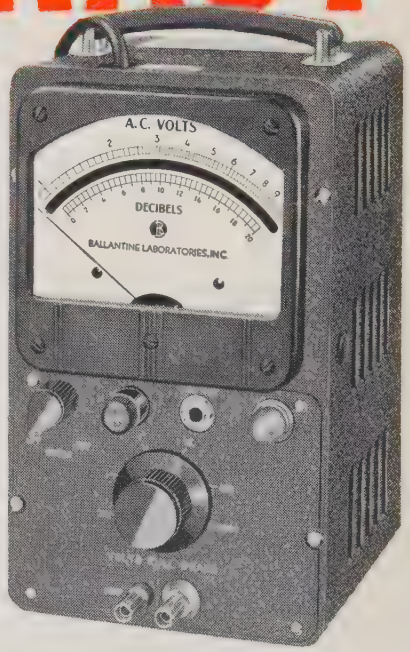
1%

ACCURACY

over

ENTIRE METER SCALE!

1 mv – 250 v, 20 cps – 20 kc



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This is the most precise instrument in our entire line of sensitive wide-band Electronic Voltmeters.

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SENSITIVE ELECTRONIC VOLTMETER

Price $315

- Top accuracy of 1% over entire meter scale from 1 mv to 250 v and over the band of 20 cps to 20 kc. Better than 2% to 1,000 volts and for the wider band of 10 cps to 250 kc.
- High input impedance: 2 megohms shunted by 15 pF, except 25 pF on lowest voltage range.
- Long life: Several thousands of hours of operation without servicing or recalibration.
- Does not require stabilized input voltage. Less than ½% change in indication with power supply change from 105 v to 125 v.
- Five inch, mirror-backed, easy-to-read meter. Only two scales with mirror between. One is 1 to 10 for volts, and the second is 0 to 20 for decibels.

  Also available in 19 inch relay rack Model 300 G-S2 at $325.

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CHECK WITH BALLANTINE FIRST FOR LABORATORY AC VACUUM TUBE VOLTMETERS, REGARDLESS OF YOUR REQUIREMENTS FOR AMPLITUDE, FREQUENCY, OR WAVEFORM. WE HAVE A LARGE LINE, WITH ADDITIONS EACH YEAR. ALSO AC/DC AND DC/AC INVERTERS, CALIBRATORS, CALIBRATED WIDE BAND AF AMPLIFIER, DIRECT-READING CAPACITANCE METER, OTHER ACCESSORIES. ASK ABOUT OUR LABORATORY VOLTAGE STANDARDS TO 1,000 MC.

FXR's COAXIAL BROADBAND BI-DIRECTIONAL COUPLERS

Model No. N616D

OPERATION: Full two octave frequency range

FREQUENCY RANGE: Model 616 — 0.25 KMc to 1.0 KMc
Model 617 — 1.0 KMc to 4.0 KMc

COUPLING: 20 ±2 db including frequency sensitivity

DIRECTIVITY: 20 db min

Individual calibration curves attached

No. 4 of a series of FXR's new precision microwave components designed to meet the ever-growing needs of the microwave industry.

FXR's Coaxial Broadband Bi-Directional Couplers are versatile components with high directivity and uniform coupling over a range of two octaves. These couplers can be employed in reflectometer setups, to monitor power and frequency or as a standard of attenuation. The built-in loads on the line can absorb 1 watt of CW power which makes the couplers ideal for the injection of large values of local oscillator power into a mixer circuit.

The main line VSWR is less than 1.2 and the auxiliary arm VSWR is less than 1.25 including the termination.

MODEL NO.	FREQUENCY RANGE KMc	CONNECTOR TYPE	PRICE
N616D	0.250-1	N Jack	$160.00
N617D	1-4.0	UG-23C/U equivalent	150.00

Calibration curves for the incident and reflected couplers are attached to each instrument.

Write for Catalog Sheet No. 616

FXR a division of Amphenol-Borg Electronics Corp.
25-26 50th Street / RA. 1-9000
Woodside, N. Y. / TWX: NY 43745

IRE People

(Continued from page 77A)

In his prior capacity as Reliability Director for Hoffman Electronics Corp., Dr. Romig worked with Dr. A. C. Floyd on the TACAN Reliability Program including the first application to a System of the AGREE Reliability Tests, these results being presented in a joint paper at WESCON in 1959. He established the Reliability Program for the AN/ULD-1 Project and aided the Semicondictor Division in establishing their Quality Engineering program.

Dr. Romig left Bell Telephone Laboratories in 1951 to go to the Hughes Aircraft Company, where he established their Quality Control System on a firm basis, operated their Central QC System and then aided in organizing autonomous systems in each Division. Prior to joining Hoffman, he served as consultant in Quality Assurance, Reliability, and Industrial Engineering for a number of firms including Firestone, Adel, Bendix-Pacific, Helipot, Telemeter Magnetics, Northrop, Gladding McBean, and Microdot.

He has been active in a large number of technical societies both locally and nationally, serving this year as Conference Chairman for the Western Regional ASQC Conference at Disneyland. He is a founding member and Fellow of ASQC, and served as National Director for four years and National Executive Director for two years, besides having been Chairman of the Los Angeles Section and also earlier the Metropolitan New York Section. He is a Fellow of the American Statistical Association, the AAAS, and a member of six Professional Groups of the IRE, serving as Vice President of PGRQC for one year.

He is also a member of the Society on Non-Destructive Testing, the Society of Applied Industrial Engineers, the American Society for Testing Materials, Econometric Society, the Institute of Mathematical Statistics (of which he was a Council Member for five years), the Biometric Society, the Institute of Management Science, the College of Planning, Sigma Xi and Alpha Pi Mu fraternities, and the American Legion. He is a New York Licensed Professional Engineer.

❖

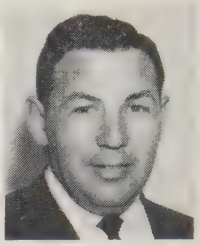
Bernard Rosen (A'51 M'56) has been appointed to the position of General Manager, Equipment Engineering, of Polarad Electronics Corporation, Long Island City, N. Y. He will direct an engineering and pilot production group with the responsibility of developing newly designed prototypes for production practicability. He joined the company in 1951.

B. Rosen

Mr. Rosen received his BEE degree from the City College of New York and his MEE from the Brooklyn Polytechnic Institute. He is a member of Sigma Xi, Eta Kappa Nu and Tau Beta Pi engineering honor societies, and belongs to the IRE Professional Groups on Antennas and Propagation, Microwave Theory and Techniques, and Radio Frequency Interference.

❖

Francis L. Ross (M'59) has been named Sales Manager of Micromodular Components, Division of Ling-Temco Electronics, Inc. He will direct the company's nationwide sales activities. Micromodular—Ling-Temco's newest division—specializes in design and production of encapsulated welded modules, including high-voltage rectifiers and also environmental qualification testing.



F. L. Ross

Mr. Ross brings extensive experience in engineering and executive sales to the position. He was a sales engineer for Sylvania Electric Products Inc., and worked as a product sales engineer at the Semicondictor Division, Hughes Aircraft Company. He also completed Pacific Telephone & Telegraph Company's engineer executive training program.

He holds the B.S. degree in engineering from U.C.L.A. and also did graduate work at that University. He served as an officer in the Marine Air Corps.

❖

Dr. Morris H. Shamos (M'54), New York University Professor of Physics, has been elected to the Board of Directors of Elm Instrument Corp., Hempstead, L. I., New York, electronics firm.

M. H. Shamos

A specialist in electronics and nuclear physics, Dr. Shamos is Chairman of the Department of Physics and Director of the Cosmic Ray Laboratory at Washington Square College of New York University. He is a consultant to the Atomic Energy Commission, the National Broadcasting Co., Tung-Sol Electric Co., and is a director of National Radiac, Inc.

Dr. Shamos is a Fellow of the New York Academy of Sciences and a member of the American Physical Society, and the American Association for the Advancement of Science. He is the author of three technical books and a number of professional journal articles and has been issued several patents.

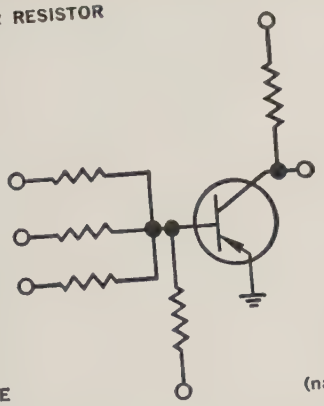
❖

William Sichak (M'46–SM'56) former Director of Communications Systems with International Telephone and Telegraph

(Continued on page 80A)

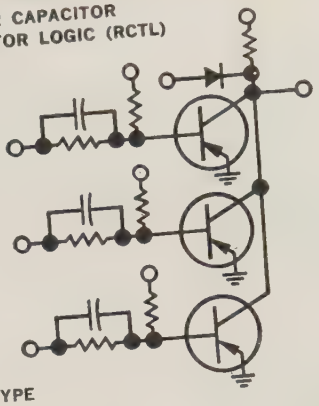
ACHIEVE STAGE DELAYS AS LOW AS 1 NANOSECOND...CLOCK RATES FROM 1 TO 1,000 Mc WITH RCA'S BROAD LINE OF COMPUTER SWITCHING DEVICES

TRANSISTOR RESISTOR LOGIC (TRL)



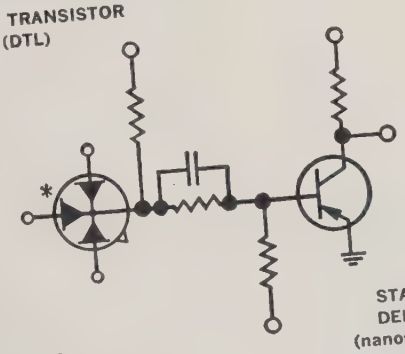
RCA TYPE	STAGE DELAY (nanosec)
2N585 (NPN ALLOY)	1000
2N404 (PNP ALLOY)	800
2N1301, 2N795 (PNP MESA)	200
2N697 (NPN SILICON)	175
2N705 (PNP MESA)	100
2N706 (NPN SILICON)	100
TA-1882 (NPN UHF MESA)	60

RESISTOR CAPACITOR TRANSISTOR LOGIC (RCTL)



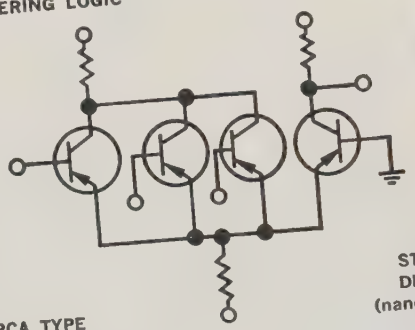
RCA TYPE	STAGE DELAY (nanosec)
2N585 (NPN ALLOY)	200
2N404 (PNP ALLOY)	160
2N1301, 2N795 (PNP MESA)	40
2N697 (NPN SILICON)	35
2N705 (PNP MESA)	13
2N706 (NPN SILICON)	22
TA-1882 (NPN UHF MESA)	9

DIODE TRANSISTOR LOGIC (DTL)



RCA TYPE	STAGE DELAY (nanosec)
2N585 (NPN ALLOY)	240
2N404 (PNP ALLOY)	180
2N1301, 2N795 (PNP MESA)	50
2N644, 2N1450 (PNP DRIFT-FIELD)	— —
2N697 (NPN SILICON)	40
2N705 (PNP MESA)	16
2N706 (NPN SILICON)	27
TA-1882 (NPN UHF MESA)	12

TRANSISTOR CURRENT STEERING LOGIC



RCA TYPE	STAGE DELAY (nanosec)
2N585 (NPN ALLOY)	120
2N404 (PNP ALLOY)	100
2N1301, 2N795 (PNP MESA)	30
2N644, 2N1450 (PNP DRIFT-FIELD)	17
2N697 (NPN SILICON)	25
2N705 (PNP MESA)	12
2N706 (NPN SILICON)	15
TA-1882 (NPN UHF MESA)	7

1N3128 to 1N3130, RCA Tunnel Diodes—Stage Delays in Tunnel Diode Logic Circuit of 5 nanosec. to 1 nanosec.

*3DG001 (Multiple-Diode Switch)

Now, over 50 RCA silicon and germanium switching devices provide comprehensive coverage of your logic-circuit needs for military and industrial data-processing equipment.

This broad spectrum of RCA switching transistors and diodes can bring flexibility and economy to your new logic-circuit designs. In the practical working circuits shown, these devices can help you meet and exceed many of the exacting speed requirements of today's military and industrial computers.

Here are some of the outstanding features of these versatile switching devices:

- RCA 3907/2N404...*Extra rugged and reliable*...Produced under RCA's Certified-Reliability Premium Switching-Transistor Program.
- RCA 2N585...Germanium NPN counterpart of the 2N404.
- RCA 2N1301...High-speed germanium switching mesa transistor in TO-5 package...minimum beta of 40, pulse repetition rates up to 10 Mc.
- RCA 2N697...High-speed, high-current silicon *planar* transistor.
- RCA 2N705...Very high-speed germanium switching mesa type.
- RCA 2N706...Very high-speed silicon *planar* switching transistor designed to meet MIL specifications.
- RCA Developmental Type TA-1882...*World's fastest* germanium mesa transistor.
- NEW RCA 2DG001, 3DG001...Multiple arrays of two and three diffused-junction germanium computer switching diodes in TO-33 package size. Typical recovery time 160 nanoseconds.

Call your RCA representative today for further information. For additional technical data, write RCA Semiconductor and Materials Division, Commercial Engineering, Section J-35-NN, Somerville, N.J.

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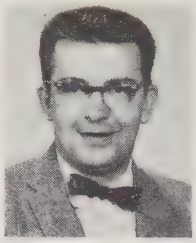
RADIO CORPORATION OF AMERICA

IRE People

(Continued from page 78A)

Federal Laboratories, has formed a new firm, Sichak Associates, East Orange, N. J., which will serve industry and the government in the fields of communications, data processing, computer technology and other areas of electronics.



W. Sichak

He is a Fellow of the American Association for the Advancement of Science. A distinguished inventor (he holds many issued patents and has others pending), he has been with ITT since 1945, and recently resigned his post.

The new laboratory intends to participate in the rapidly expanding fields of space communications, exploration, probes and other scientific efforts, performing studies and developing equipment. A significant portion of the firm's efforts, according to Mr. Sichak, its President and Chairman of the Board, will be devoted to basic studies in the physical sciences.

Mr. Sichak, a native of Ambridge, Pa., and a graduate of Allegheny College, Meadville, Pa., with the Bachelor's degree in physics, served for three years at the Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, before joining ITT. He is a leader and internationally recognized authority in O/H (over-the-horizon) systems for long-distance telephone and television. He played a major role in developing these systems and has extensive experience in the fields of antennas and microwaves.

He has authored many technical papers which have been published by the IRE, and has contributed to *Electronics*, *Electrical Communication*, and others. He is a member of the American Physical Society, New York Academy of Sciences and the American Rocket Society. He has served on committees of the IRE and the Electronic Industries Association, and presently is active in reviewing papers for various IRE Professional Groups and the *Review of Scientific Instruments*.

❖

Mark Shephard, Jr. (SM'53) has been appointed Executive Vice President of Texas Instruments, Inc., Dallas, with responsibility for all company operating activities.

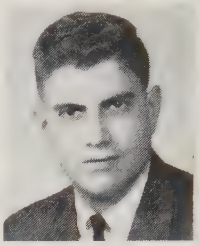
He joined the company in 1948. He has been in charge of all semiconductor activities since Texas Instruments' entrance into that field in 1952, and Vice President in charge of the former Semiconductor-Components Division since 1955.

He received the Bachelor's degree in electrical engineering at Southern Methodist University, Dallas, Tex., in 1942, and the Master's degree in electrical engineering at the University of Illinois, Urbana, in 1947. During World War II Naval Service he was a Lieutenant (JG) on a cruiser as a specialist in radar and electronics maintenance.

Mr. Shepherd is a member of the Armed Forces Communications & Electronics Association.

❖

David M. Smith (M'59) has joined the East Coast office of the Univac Military Applications Department, it was recently announced. He has had an extensive background in military systems design management, and will advise Univac's military customers on the design and development of weapons and control systems.



D. M. Smith

He was formerly Manager of the Ramo-Wooldridge office at Fort Monmouth, N. J., and had previously spent five years with the Burroughs Corporation as a design engineer, sales engineer, and manager

Mr. Smith is an electrical engineering graduate of Pennsylvania State University, University Park.

❖

The appointment of **Robert T. Smith** (M'56) to the position of Chief Engineer of the newly-formed Jensen Engineering, Inc., of Palo Alto, Calif., was recently announced.

R. T. Smith

He was formerly with the Philco Western Development Laboratories in Palo Alto, where he was Director of Satellite Control Development.

He received the B.S. degree in electrical engineering from the University of Kansas, Lawrence, in 1952. From 1952–1958, he was employed at USNOTS, China Lake, Calif., where he was responsible for development of control systems, fuzing systems for air-launched weapons, and air-to-ground fire control systems. At Philco, he had system management responsibility for communication and control sub-systems for Air Force satellite programs.

Mr. Smith is a member of Sigma Tau and Eta Kappa Nu.

❖

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IRE DIRECTORY

Babcock Electronics Corporation, Costa Mesa, Calif., has appointed **W. W. Smith** (A'45) vice president, engineering. He was chief of engineering development prior to assuming his new post. Before coming to Babcock, he was associated with the Gonset Division of Young Spring and Wire Corporation.

W. W. Smith

Smith is the author of a number of technical books on electronic subjects and has published nearly 100 technical articles in national electronic magazines. In addition, he is credited with the original development of pre-modulation speech clipping and filtering techniques.

Prior to World War II, he was editor of *Radio Magazine* and the *Radio Handbook* and subsequently became editorial director and chief engineer of Editors and Engineers Limited, Technical Publishers and Consultants. During the early phases of World War II, he was the top civilian radar engineer for the Marine Corps and was largely responsible for the initial development of AN/TPS-1 radar equipment.

Smith was awarded the Commendation Ribbon and a citation by the Secretary of the Navy both for his work prior to entering the service and for his accomplishments while in uniform, all relating to contributions toward the development or improvements of radar and various types of communications for the U. S. Navy and U. S. Marine Corps. He subsequently served as a civilian consultant to the Amphibious Branch of the Office of Naval Research.

❖

Paul Steigman (M'59) has been appointed Vice President in Charge of Technical Services for Standard Systems Corporation.

P. Steigman

Prior to joining Standard Systems, he was employed by Executone, Inc. as Director, Technical Information. He has also worked for Bell Telephone Laboratories and General Electric. He is a graduate engineer of Pratt Institute of Technology and has attended graduate school at the Polytechnic Institute of Brooklyn.

❖

The board of directors of Mikros, Inc., newly formed subsidiary of Electro Scientific Industries, Portland, Ore., has elected **Douglas C. Strain** (S'47-A'49-M'54-SM'60), president; **Thomas J. Holce** (M'60) and **Paul D. Lintner** (M'61), vice presidents; and **Laurence A. Morin** (S'49-A'52-M'57), secretary-treasurer.

Morin, now Mikros general manager, was formerly sales manager for the parent company. Lintner, marketing manager for

(Continued on page 82A)

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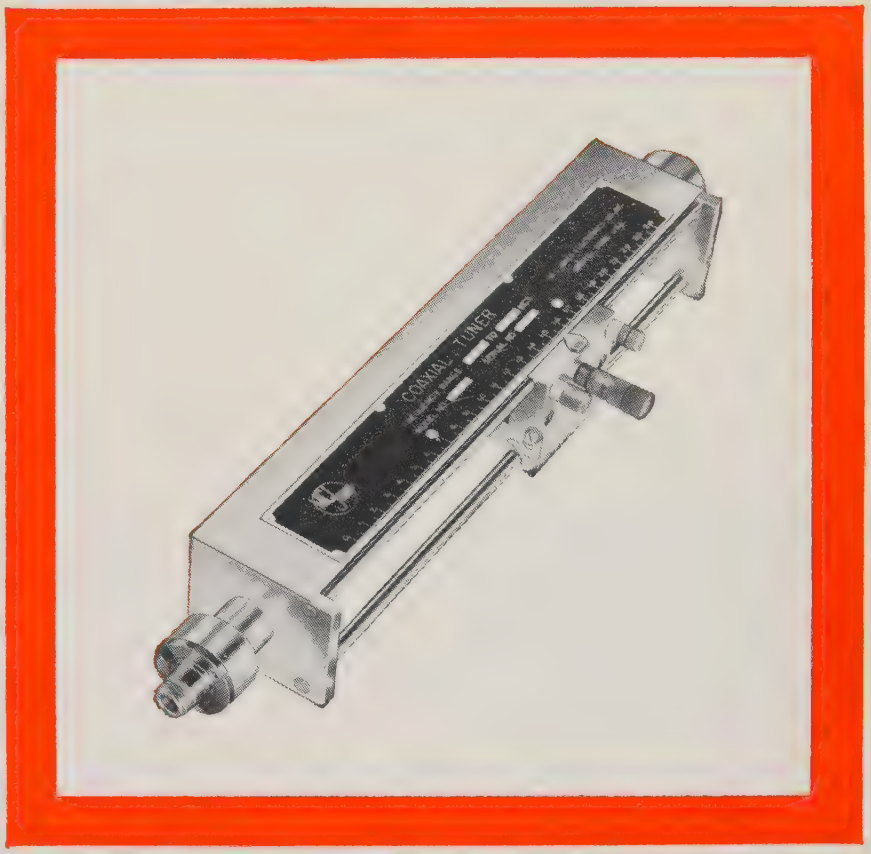

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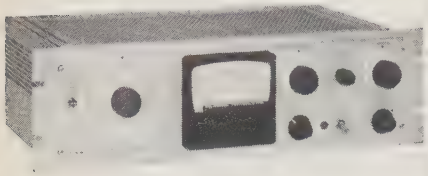
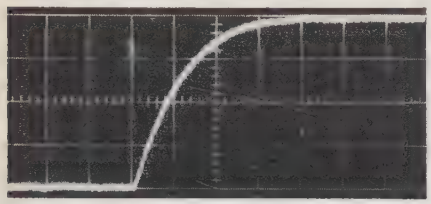


The new Keithley Model 415 micro-microammeter offers high speed of response, accuracy, and zero suppression.

A speed of response of less than 600 milliseconds to 90% of final value at 10^{-12} ampere is possible where external circuit capacity is 50$\mu\mu$f. Accuracy is ±2% of full scale on 10^{-3} through 10^{-8} ranges and ±3% on ranges below. Zero suppression permits full scale display of one per cent variations of a signal.

The 415 is ideal for use with ion chambers, ionization gages, gas chromatography, mass spectrometry.



Response to a current step of 10^{-12} amp. Input capacity is 35 $\mu\mu$f. One major horizontal division equals 200 milliseconds.

SPECIFICATIONS

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IRE People

(Continued from page 81A)

the new firm, was formerly ESI's sales manager in charge of ESIAC Computers. Holce, Mikros chief engineer, served as project engineer for ESI on electron microscope development. Strain is president of both companies.

The subsidiary was formed to manufacture a new electron microscope and associated instruments including a high voltage power supply, a high vacuum evaporator and the Mikros cardiac defibrillator.

❖

The appointment of **Elmer Thomas** (M'58) as Assistant Director of Research and Development at Page Communications Engineers, Inc., has been announced.

He has been Manager of the Page Research and Development Laboratory since 1957. He is the inventor of the Decision Threshold Computer, an important Page development which permits increased orders of diversity in the presence of selective fading, and he has applied the invention to high-frequency and very-high-frequency scatter circuits with marked success.

Prior to joining Page, he was employed by the Department of Defense for two years as head of the Time-Division Multiplex Section, responsible for the design and development of new types of multiplex equipment.

Mr. Thomas received the B.S.E.E. degree from the University of Pittsburgh, Pa.

❖

Dr. James A. Van Allen (SM'59), the physicist who in 1958 led in the discovery of the radiation belts around the earth which now bear his name, has been named an Elliott Cresson Medalist of The Franklin Institute.

J. A. Van Allen

Dr. Van Allen, Chairman of the Department of Physics and Astronomy at the State University of Iowa and a top-level advisor to the National Aeronautics and Space Administration, will receive his medal at the Institute's annual "Medal Day" ceremonies October 18.

The radiation belts were identified by Van Allen from data or the lack of it received from SUI-built instruments aboard Explorer I, the first U. S. Satellite and succeeding Explorers and Pioneer deep space probes. In low altitude areas a Geiger counter with an electronic setup indicated radiation counts of an intensity which agreed fairly well with theoretical extensions of earlier rocket experiments. However, at high altitude, the Geiger counter very unexpectedly sent back no reports. He and his associates reasoned that it was more likely that the counter ran into radiation so intense that it could not cope with the counting, rather than the reverse situation of no radiation at high altitudes.

He proved his theory in the laboratory by subjecting a duplicate instrument package to a powerful beam of X-rays which caused the counter to go temporarily dead when overloaded. Later satellite experiments using Geiger counters of greater dynamic range verified the presence of the Van Allen radiation belts.

He is credited with furthering space research since 1945 through the development of instrumentation and measurement techniques. He supervised both the development of the Aerobee rocket for upper atmospheric probing and the "rockoon" combination in which a high altitude balloon is used as the launching platform for a rocket.

It was during the course of the rockoon research that Dr. Van Allen and others came up with the idea of an international scientific effort in geophysics. This was the genesis of the recent International Geophysical Year.

The medal citation to Dr. Van Allen reads: "In consideration of his many contributions and pioneering achievements in the field of space science; in particular, for his discovery of regions of trapped radiation in space, now termed the 'Van Allen Radiation Belts.'"

James Alfred Van Allen was born Sept. 7, 1914, in Mount Pleasant, Iowa. He received his Bachelor of Science degree in Physics at Iowa Wesleyan College in Mount Pleasant and his Master of Science and doctoral degrees at the State University of Iowa, Iowa City. For the next two years, Dr. Van Allen was a research fellow at Carnegie Institution in Washington. During the Second World War, the physicist served with the U. S. Navy, primarily stationed at the Bureau of Ordnance to work on development of proximity fuses. In 1946, he joined the staff of the Applied Physics Laboratory at Johns Hopkins University as head of high altitude research. Five years later he returned to the State University of Iowa as Chairman of the Department of Physics. In 1949 Dr. Van Allen was awarded the Hichman medal of the American Rocket Society for his development of the Aerobee rocket, and the Physics Award of the Washington Academy of Science.

He is a Fellow of the American Physical Society and American Rocket Society, and a member of the American Geophysical Union, National Academy of Sciences, International Academy of Astronautics, American Astronautical Society, Sigma Xi, and other scientific groups.

❖

Norman Weinhouse (A'54–SM'60) has been named chief engineer of Kearfott Microwave Division, General Precision, of Van Nuys, California. Since joining the firm in 1958, he has served as manager of the Systems Department, and as manager of the Engineering Technical Office. In

(Continued on page 84A)

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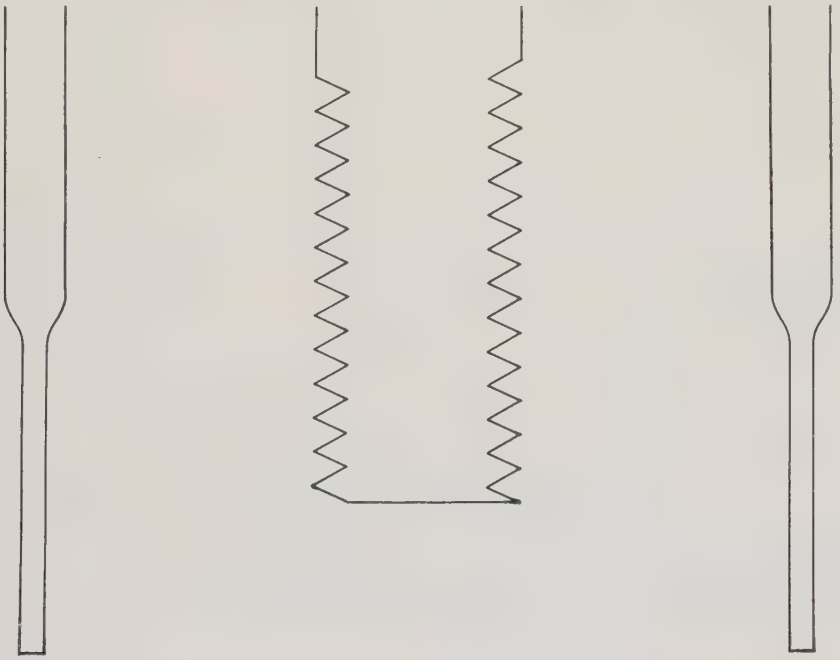

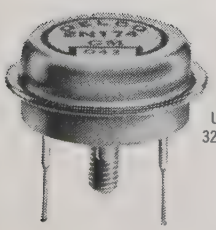

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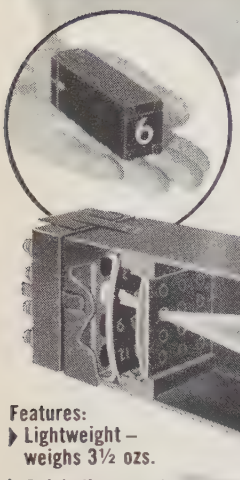

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(Continued from page 82A)

these positions he was responsible for the design of systems utilizing the microwave art, including radar test sets, monopulse radar systems, and microwave systems.

Prior to joining Kearfott, Mr. Weinhouse worked with Collins Radio for six years, leaving as manager of their Systems Engineering Group.

✣

Dr. Lester C. Van Atta (M'42–SM'43–F'52) has been appointed technical director of Hughes Aircraft Company's research laboratories at Malibu, Calif., it was announced recently.

L. C. Van Atta

Dr. Van Atta has just completed a year's service in Washington, D. C., as special assistant for arms control for the director of defense research and engineering in the office of the Secretary of Defense. He was on professional leave of absence from Hughes during his government service.

Dr. Van Atta joined Hughes in 1950 as head of the company's microwave laboratory, and later became associate director of the research laboratories, which then were located at the Culver City, Calif., headquarters.

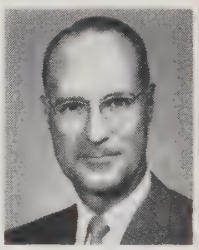
For three years preceding his government service, he was director of technical information and education, starting or implementing many of the company's fellowship and aid-to-education programs. He was one of the leaders in founding the Southern California Industry Education Council and served as the council's vice president in 1959–60.

Dr. Van Atta received a bachelor's degree from Reed College, Portland, Ore. in 1927 and a Ph.D. degree from Washington University, St. Louis, Mo. in 1931. Before joining Hughes he headed the antenna research branch at the Naval Research Laboratories.

He is a Fellow of the Institute of Radio Engineers, where he is presently a member of the board of directors and previously served as vice chairman of the committee on professional groups. He is also a Fellow of the American Physical Society, the American Association for Advancement of Science, and the American Geophysical Union.

✣

J. H. Cannon, Vice President of Cannon Electric Company, Los Angeles, has announced the appointment of **Dr. Edward G. Witting** (SM'56) as Manager of Research and New Products. He leaves the post of Deputy Assistant Secretary of the Army (Research and Development) to assume the new research position at Cannon Electric.



E. G. Witting

He attended Johns Hopkins University, Baltimore, Md., received the Ph.D. degree in physics in 1940, and remained to conduct extensive research. He was later employed by the Army Signal Corps Laboratories and worked on the development of underwater sound detecting equipment. During military service, he served as Executive Officer of Squier Signal Laboratory, and also as Army Liaison Officer to the Naval Research Laboratory.

In 1946, he returned as a civilian to the Signal Corps Engineering Division Laboratory. During the following ten years, he was employed as Chief of the Engineering Staff, as Director of the Physical Sciences Division, and finally as Chief Scientist of the Signal Corps.

In June, 1956, he was appointed Deputy Director of Research and Development, Department of the Army, and, in March, 1961, Deputy Assistant Secretary of the Army (Research and Development).

Dr. Witting is a member of the Acoustical Society of America, the American Physics Society, the Armed Forces Communications and Electronics Association, and has served as Chairman of the IRE Professional Group on Military Electronics.

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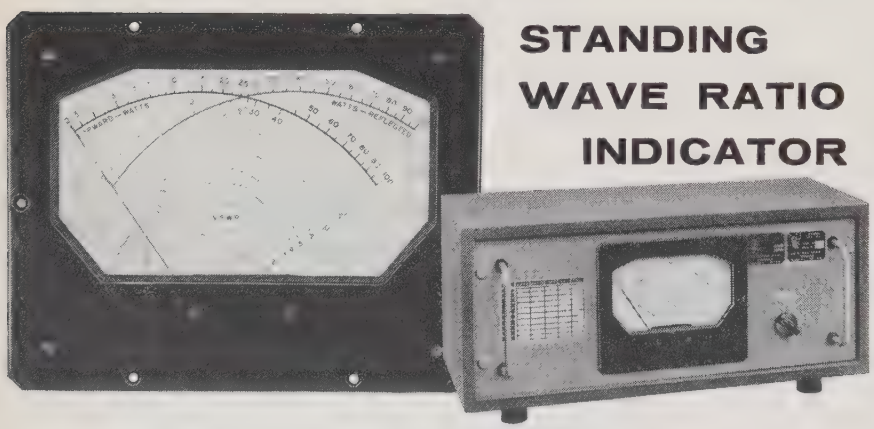
(Continued on page 86A)

SWR-3000



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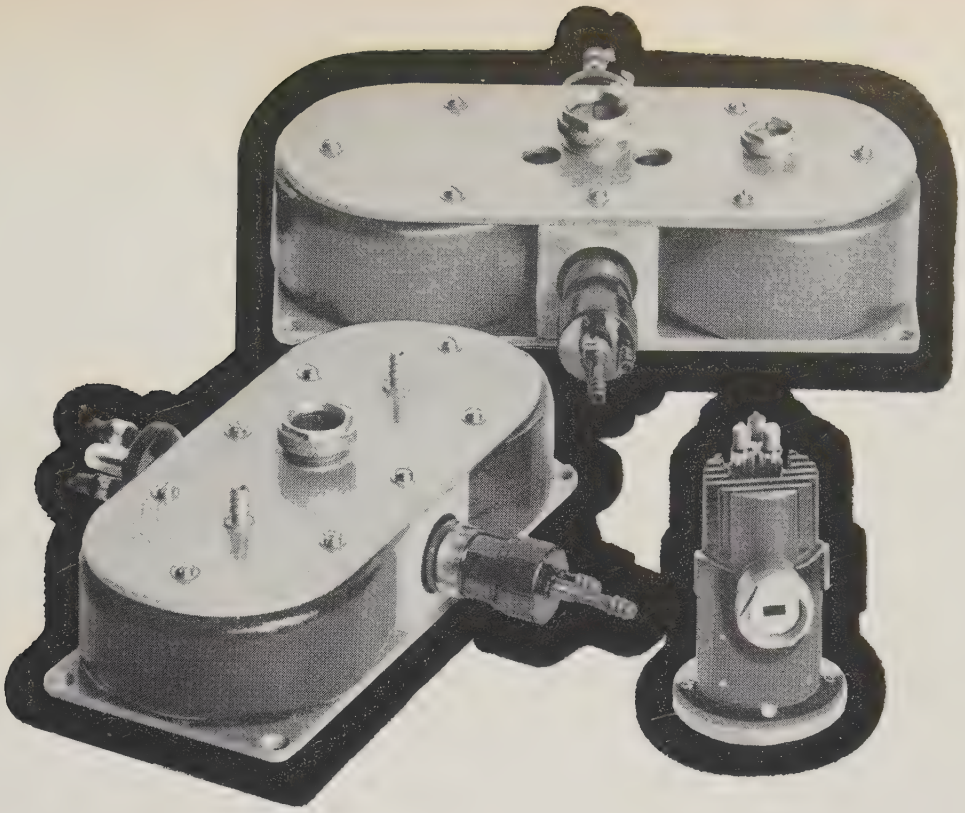

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L-3691	68 - 80	1500	0.100 average	Liquid
L-3640	48 - 52	Fixed	1.0	Liquid
L-3693	48 - 52	1500	1.0	Liquid
L-3629	33 - 37	Fixed	15.0	Liquid
L-3628	33 - 37	1600	10.0	Liquid
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● **Aerospace Telemetry***

Harry L. Stiltz, Lockheed (Ga.), Editor

Explains purpose of telemetry, how used, basic terminology, most widely employed techniques, how to choose proper system, how to set up measurements program, type of equipment required, methods of cooling equipment, pre-detection recording, high speed FAM/FM and PCM/FM, sampling theory and interpolation. All telemetry techniques covered by latest (Dec., 1960) IRIG standards. Each chapter prepared by specialist (of Lockheed, G-E, Ampex, etc.,) on a particular phase of telemetry. **Sept., 1961—505 pages—6 × 9″—illus.—$15**

● **Reliability: Theory & Practice***

Igor Bazovsky, United Control Corporation

Presents, step by step, the important phases of a reliability program: how to eliminate early failures, how to prevent wearout failures, how to prevent, reduce or eliminate chance failures. Shows how newest reliability methods can be properly implemented. Identifies the seven tools of reliable design. Recommends minimum five-point reliability portion of a quality and reliability control program for production. Helps get maximum reliability in least time for minimum cost. **Aug., 1961—296 pages—6 × 9″—illus.—$10.95**

● **Electronic Computers**

Paul Von Handel, Institute for Defense Analysis, et al.

Surveys the basic principles and applications of computers for engineers and scientists who want to know the performance of computers, as well as for executives who are mainly concerned with their economical aspects. Written by distinguished experts in the field of computers, stressing principles and methods rather than engineering detail. **Aug., 1961—236 pages—6 × 9″—$13.50**

● **Semiconductor-Diode Parametric Amplifiers**

Lawrence A. Blackwell, Texas Instruments & Kenneth L. Kotzebue, Watkins-Johnson Co.

May, 1961—200 pages—6 × 9″—illus.—$9

● **Principles of Electronic Warfare***

Robert J. Schlesinger, et al., of Convair

Aug., 1961—224 pages—6 × 9″—illus.—$8

● **Handbook of Laplace Transformation**

Floyd E. Nixon, The Martin Co.

Oct., 1960—115 pages—5⅜ × 8″—$4.95

● **Proceedings of the National Symposium on Machine Translation**

H. P. Edmundson, Planning Research Corp., et al.

June, 1961—544 pages—6 × 9″—illus.—$7.95

● **Computer Programming Handbook: Guide for Beginners**

Robert Nathan, Univ. of Calif. at Berkeley & Elizabeth Hanes, Calif. Inst. of Tech.

Oct., 1961—224 pages—6 × 9″—illus.—$7.65

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(Continued from page 84A)

Directors of Hazeltine Corporation have announced the election of **Webster H. Wilson** (M'47) as Chairman of the Board. Mr. Wilson, who is also president and chief executive officer of the electronics firm, succeeds W. A. MacDonald, who passed away recently.

The chairman and president joined Hazeltine in 1946 and was elected a director in 1958 and president in 1960. He has been a chief project administrator, vice president and later executive vice president for operations of Hazeltine Electronics Division. He received his B.S. degree from Massachusetts Institute of Technology in business administration and aeronautical engineering. As a naval officer during World War II, he supervised the installation and maintenance of shipboard electronic equipment.

❖

Hilmer J. Wissemann (A'46–SM'52) has been appointed Vice President in charge of the Government Products Group in the Apparatus Division of Texas Instruments, Inc., Dallas, Tex.

He joined the company in 1945 in the Laboratory and Manufacturing Division. He was elected Assistant Vice President in charge of Engineering for the Apparatus Division in 1954, and was placed in charge of the Government Products Group when it was formed in 1961.

Following his graduation from Southern Methodist University, Dallas, Tex., Mr. Wissemann was a high school physics teacher for two years. From 1941–1945 he served as an officer in the Navy, specializing in the use, repair and supervision of electronic equipment.

❖

Membership

The following transfers and admissions have been approved and are now effective:

Transfer to Senior Member

Baruch, J. J., Cambridge, Mass.
Bowness, C., Weston, Mass.
Carlson, E. W., La Jolla, Calif.
Granger, F. P., Jr., Santa Barbara, Calif.
Joyal, H. J., Manchester, N. H.

Admission to Senior Member

Hill, W. S., Weston, Conn.
Hipple, G. M., Columbus, Ohio
Kittinger, W. T., Houston, Tex.
Klein, C. A., Lexington, Mass.
Lesgold, R. B., Milwaukee, Wis.
Mahn, G. R., Hales Corners, Wis.
Middendorf, W. H., Covington, Ky.
Radin, E. L., Arlington, Va.
Rodriguez, M. J., Massapequa Park, L. I., N. Y.
Stroble, R. L., Milwaukee, Wis.
Vander Tuin, J. A., Sunnyvale, Calif.

Transfer to Member

Aurora, J. R., Granby, P. Q., Canada
Bowen, D. E., Richardson, Tex.,
Eisele, E. J., Brookfield, Wis.
Ellert, F. J., Schenectady, N. Y.
Fields, W. H., E. Setauket, L. I., N. Y.
Fields, W. H., E. Setauket, L. I., N. Y.
Hickey, L. F., Derby Kans.
Ishee R. G. Gulfport Miss.
Kear F. W., Albuquerque, N. M.
Norlander, F. W., Union City, N. J.
Ramani, R. V., Coimbatore, Madras, India
Schwartz, R. N., Williamsville, N. Y.
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(Continued on page 89A)

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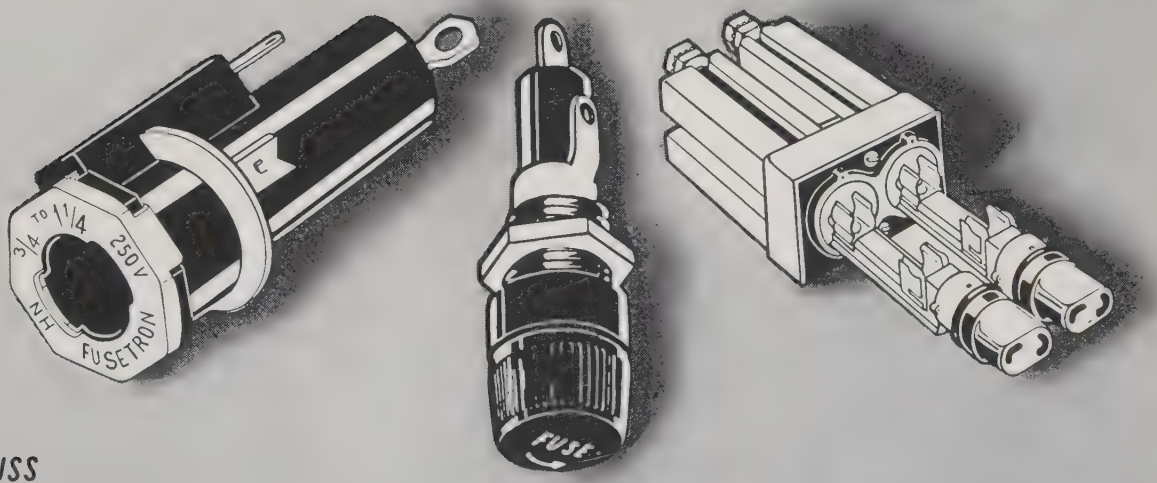

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Get full data for your files, write for BUSS bulletin, Form SFB—it gives a comprehensive picture of the complete line of fuses and fuse mountings of unquestioned high quality.

BUSSMANN MFG. DIVISION, *McGraw-Edison Co.,* **UNIVERSITY AT JEFFERSON, ST. LOUIS 7, MO.**

An NEC semiconductor device for every circuit application

The range of semiconductor products at NEC is perhaps the widest of any manufacturer. Entertainment or industrial, standard broadcast frequencies to microwave — NEC has a semiconductor device with the ratings and characteristics for your application. There's a wide selection of current ratings, operating temperatures, and mountings.

These are produced in NEC's new semiconductor plant where crystal surfaces are cleaned by 100 tons of hyper pure water daily. Every seal is tested in krypton isotopes, providing a failure rate suitable for the most critical applications.

A single source for semiconductors can mean a saving in time and costs. Just let NEC know your requirements and full technical data will be sent.

Types of NEC semiconductor devices

• PNP super-grown Germanium transistors • Germanium photo transistors • PNP alloy junction Germanium transistors • Silicon Mesa transistors • Germanium Mesa transistors • Germanium gold-bonded diodes • Silicon rectifiers • Silicon controlled rectifiers • Silicon capacitors • Zener diodes • Germanium point-contact mini-diodes • Silver-bonded diodes • Microwave mixer diodes • Silicon junction mini-diodes

Reliability The first NEC transistorized carrier telephone system was an NT & T installation in 1958 between Toyama and Takaoka, a distance of 15 miles. It consists of two terminal stations and a repeater station with 240 channels using 1,600 transistors.
During last 14,000 hours of operation transistor failure has caused only two channel faults. This corresponds to a failure rate of 0.009% per 1,000 hours.

Communications Systems / Electronic Components NEC *Nippon Electric Co., Ltd.* Tokyo, Japan

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(Continued from page 86A)

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Thompson, J. F., Minneapolis, Minn.
Tillotson, A. D., Haddonfield, N. J.
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Turnbull, J. L., Peekskill, N. Y.
Vaughan, D. P., San Gabriel, Calif.
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Waller, P. S., Van Nuys, Calif.
Warren, R. F., Haddonfield, N. J.
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Welti, C. R., Los Angeles, Calif.
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Wintermute, E. H., IV, State College, Pa.
Winton, H. M., Huntington, L. I., N. Y.
Wolf, Robert B., Downey, Calif.
Wong, N. S., Los Angeles, Calif.
Woodbridge, R. G., 3rd, Princeton Junction, N. J.
Wyllys, R. E., Chatsworth, Calif.
York, R. B., Jr., Huntsville, Ala.
Yorks, R. G., Ann Arbor, Mich.
Zarhi, Y., Tel-Aviv, Israel
Zolnay, S. L., Columbus, Ohio

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Altomare, S. J., Brooklyn, N. Y.
Asher, R. B., Daly City, Calif.
Baumann, R. H., Dayton, Ohio
Booher, R. B., Perrysburg, Ohio
Bowen, D. E., Richardson, Tex.
Bradley, F. W., Utica, N. Y.
Broughton, J. K., Winston-Salem, N. C.
Brown, S., Los Angeles, Calif.
Cloud, E. L., Bellefontaine, Ohio
Coburn, A. R., Dedham, Mass.
Cochran, J. R., Baltimore, Md.
Conte, E. A., Rego Park, N. Y.
Cottrell, E. P., Elbridge, N. Y.
Daniels, B. T., Doylestown, Pa.
Daror, E. S., New York, N. Y.
Davis, E. R., New York, N. Y.
Dawes, A. J., Los Altos, Calif.
De Lashmitt, C. M., Thomasville, Ala.
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Edmondson, R. O., Biloxi, Miss.
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Gerstenfeld, A., Brooklyn, N. Y.
Gillett, G. M., Seattle, Wash.
Fantarella, J. R., Clinton, Conn.
Hagerthey, R. D., Northfield, N. J.
Hailman, H. F., Kalamazoo, Mich.
Halkin, H. R., Palo Alto, Calif.
Hamic, J. H., Jr., Temple City, Calif.
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Ishman, C. E., Chicago, Ill.
Johnsen, K., Hanover, N. J.
Jube, E. H., New York, N. Y.
Keipert, F. A., Alamogordo, N. Mex.
Khan, A. S., Suva, Fiji Islands
Kidder, N. R., Cleveland, Ohio
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Knight, C. H., Jr., San Diego, Calif.
Konishi, Y., Tokyo, Japan

(Continued on page 90A)

PRECISE, RELIABLE POWER SUPPLIES IN A WIDE CHOICE OF OUTPUT RANGES

SM GROUP

Optional 0.1% or 0.01% regulation:

Three rack sizes: 8¾" H, 5¼" H, and 3½" H. Impervious to operational damage: circuit protection is an inherent function of input transformer and regulator characteristics.

3½" PANEL HEIGHT

0.1% REGULATION MODELS	DC OUTPUT RANGE VOLTS	AMPS	0.01% REGULATION MODELS
SM 14-7M	0–14	0–7	SM 14-7MX
SM 36-5M	0–36	0–5	SM 36-15MX
SM 75-2M	0–75	0–2	SM 75-2MX
SM 160-1M	0–160	0–1	SM 160-1MX
SM 325-0.5M	0–325	0–0.5	SM 325-0.5MX

5¼" PANEL HEIGHT

SM 14-15M	0–14	0–15	SM 14-15MX
SM 36-10M	0–36	0–10	SM 36-10MX
SM 75-5M	0–75	0–5	SM 75-5MX
SM 160-2M	0–160	0–2	SM 160-2MX
SM 325-1M	0–325	0–1	SM 325-1MX

8¾" PANEL HEIGHT

SM 14-30M	0–14	0–30	SM 14-30MX
SM 36-15M	0–36	0–15	SM 36-5MX
SM 75-8M	0–75	0–8	SM 75-8MX
SM 160-4M	0–160	0–4	SM 160-4MX
SM 325-2M	0–325	0–2	SM 325-2MX

FOR COMPLETE SPECIFICATIONS ON MORE THAN 175 STANDARD MODEL POWER SUPPLIES, SEND FOR KEPCO CATALOG B-611.

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ATR PRODUCTS FOR MODERN LIVING

ATR UNIVERSAL INVERTERS

A.C. Household Electricity anywhere . . . in your own Car, Boat or Plane!

- Tape Recorders • TV Sets
- Dictating Machines • Radios
- Public Address Systems
- Electric Shavers • Record Players • Food Mixers • and Emergency Lighting.

	NET
6U-RHG (6 V.) 125 to 150 W. Shp. Wt. 27 lbs.	$66.34
12U-RHG (12 V.) 150 to 175 W. Shp. Wt. 27 lbs.	$66.34

Auto Plug-in Home-type Portable

ATR BATTERY CHARGERS

NO INSTALLATION . . . PLUG INTO CIGARETTE LIGHTER RECEPTACLE!

Keeps car battery fully charged in your own garage! Needed more now than ever before—makes motor starting easy! Operates directly from standard 110 volts A.C.current.

	NET
612CA4 (4 amp.) 6/12 V. Shp. Wt. 6 lbs.	$19.46
612CA6 (6 amp.) 6 /12 V. Shp. Wt. 8 lbs.	$22.46
612CA10 (10 amp.) 6/12 V. Shp. Wt. 10 lbs.	$27.71

ATR SHAV-PAKS

Keep Clean-Shaved! Plugs into Cigarette Lighter Receptacle. Keep in Glove Compartment. Operates Standard A.C.

- ELECTRIC SHAVERS
- Small Timing Devices . . . In CARS, Buses, Trucks, Boats, or Planes.

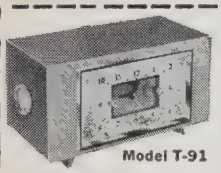
	NET
6-SPB (6 V.) 15 W. Shp. Wt. 2½ lbs.	$7.97
12-SPB (12 V.) 15 W. Shp. Wt. 2½ lbs.	$7.97

ATR ELECTRONIC TUBE PROTECTORS

Will Double or triple the life of all types of electronic tubes, including TV picture tube.

Automatic in operation, for use with any electronic equipment having input wattage of 100 to 300 watts. Fuse protected, enclosed in metal case for rugged construction and long life.

MODEL 250 (Wall Model) 115 V. A. C. Shp. Wt. 1 lb.
DEALER NET $2.63



Model T-91

MODERN TABLE ATR RADIOS

Trim, modern clock radio in ebony or ivory plastic. Powerful 5 tubes including rectifier AM radio chassis with built-in "Magna-Plate" antenna. Full-toned 4" PM speaker. Popular features include: Musical Alarm—radio turns on automatically at any pre-set time; Sleep Selector—lulls user to sleep; Automatic Appliance Timer—outlet on back of radio times any electric appliance automatically (up to 1100 watts). Cabinet 10½ in. wide, 5 in. high, 5⅛ in. deep. Wt. approx. 8 lbs.

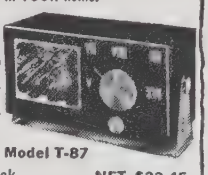
Clock Specifications:

- Genuine Telechron Movement.
- Self-Starting . . . Never Needs Winding, Oiling, or Regulating.
- Simplified Clock Controls for Radio and Sleep Switch.
- 1100 Watt Controlled Outlet.
- Automatic Buzzer Alarm.
- Gold-Plated Bezel and Numerals on Large Bone-White Dial. UL Approved.

ATR HAND WIRED—NO PRINTED CIRCUITRY "TILT-A-STAND" RADIO

MOUNT ON THE WALL—UNDER A SHELF—OR SET ON A TABLE. PERFECT for every room in YOUR home.

Power-packed 5 tubes including rectifier chassis. Built-in loop antenna. Automatic volume control. Full 4" Alnico 5 speaker. Distinctive Roman numerals on dial. Size: 9½" W x 4" D x 5¾" H. AC/DC. U.L. approved. Beautiful bakelite cabinet—Resists heat. Shipping Weight 5½ lbs.



Model T-87

	NET
Model T-91 Clock Radio, Black	$22.45
Model T-91 Clock Radio, Ivory	23.15
Model T-87 (Tilt-A-Stand), Black	17.47
Model T-87 (Tilt-A-Stand), Red	18.10

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Quality Products Since 1931
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Windom, R .N., Cocoa, Fla.
Yetter, E. J., Baltimore, Md.
Young, G. A., Plainview, L. I., N. Y.

Professional Group Meetings

Antenna and Propagation

Los Angeles—May 25

"Recent Problems of Antenna Design and Their Solutions," Prof. L. J. Chu, Mass. Inst. Tech.

Antennas and Propagation Microwave Theory and Techniques

San Diego—July 11

"Applications of Microwaves to Geodetic Surveying," E. Hose, Cubic Corp., San Diego, Calif.

Seattle—May 22

"Plasma Microwave Interactions," R. S. Elliot, UCLA, Los Angeles, Calif.

Seattle—February 23

"Ferrite Devices—Present and Future," H. Scharfman, Raytheon, Waltham, Mass.

Audio

Washington, D. C.—June 12

"The Recently Approved FM Stereophonic Broadcast System," W. H. Beaubien, General Electric Co., Utica, N. Y. and C. G. Eilers, Zenith Radio Corp., Chicago, Ill.

Washington, D. C.—November 29

"Mid-Range and Tweeter Speakers," W. A. Beaverson, Electro-Voice, Inc., Buchanan, Mich.

Bio-Medical Electronics

Los Angeles—April 13

"Infrared Spectroscopy," R. O. Brace, Beckman Instruments, Fullerton, Calif.

"Emission Spectroscopy," H. J. Sloane, Beckman Instruments, Fullerton, Calif.

Broadcasting

Omaha-Lincoln—August 4

Tour of KFAB AM and FM Transmitters.

Pittsburgh—May 14

Tour of Air Defense Missile Master Site, Major Bruckner, U. S. Army Signal Corps.

Circuit Theory

Albuquerque-Los Alamos—November 21

"Linear Transistor Theory and Measurements," Prof. H. E. Tompkins, University of New Mexico.

Electron Devices

Los Angeles—June 8

"Applications of Tunnel Diodes," K. K. Chang, RCA, Princeton, N. J.

Electronic Computers

Los Angeles—June 15

"Programming Languages and Their Implications," T. B. Steel, Jr., Systems Development Corp.

Electronic Computers Microwave Theory and Tecniqhues

Omaha-Lincoln—July 18

Tour of 789th Air Control and Warning Squadron Facilities.

"Radomes and Radars," Lt. D. Wallis, L. Welle, Communications Officers.

"FAA Microwave Relay," R. Coleman, R. Casson, D. Dougherty, FAA.

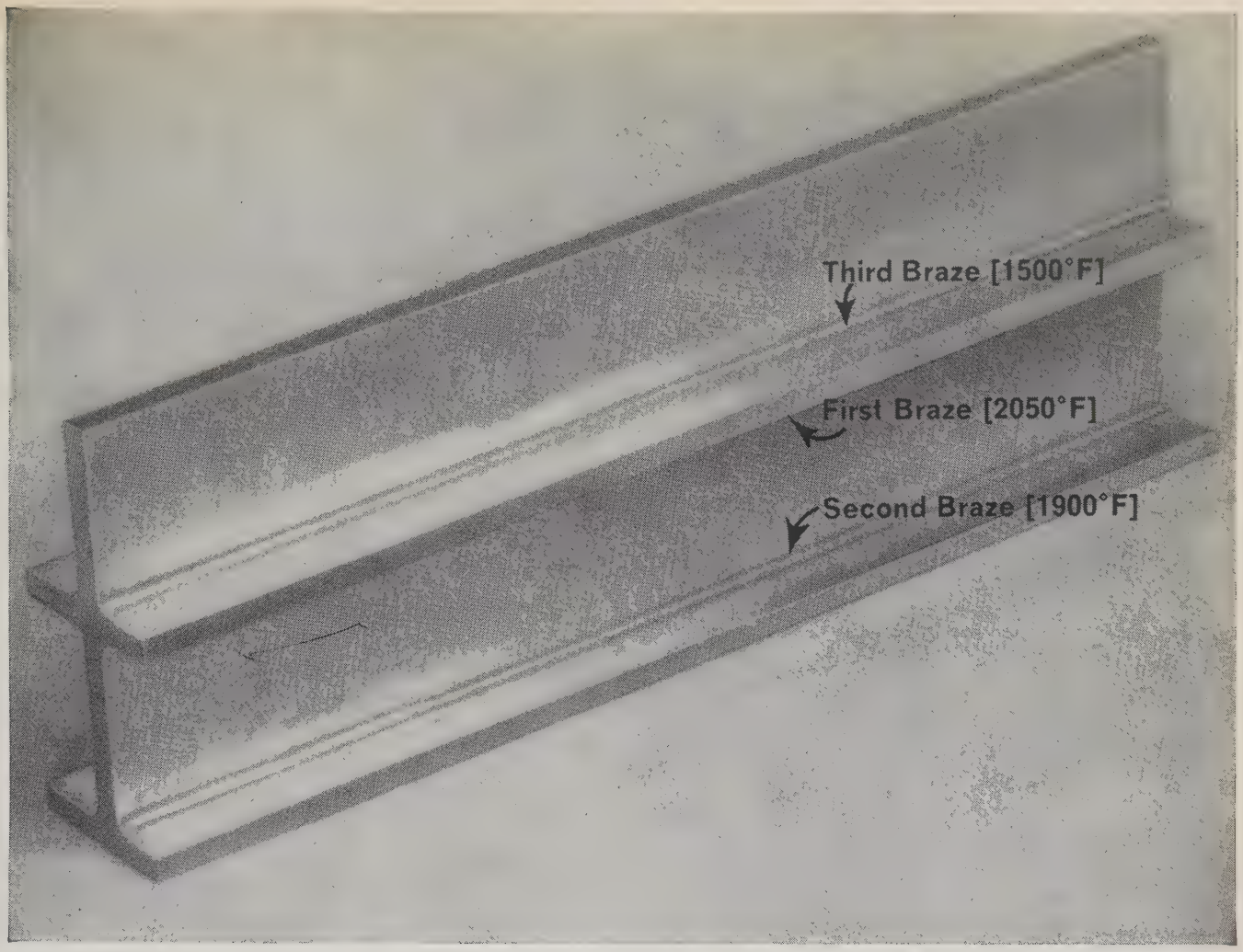
"Computer for Dataphone Digitalization," A. Wentzler, F. Schillinger, Burroughs.

Engineering Writing and Speech

Chicago—April 14

"The Element of Communication in Writing," E. L. Hilts, Armour Research Foundation, Chicago, Ill.

(Continued on page 92A)



MONEL Alloy 404 . . . a new nonmagnetic alloy that can be multiple brazed in wet hydrogen

During multiple brazing operations, furnace atmospheres must contain enough water vapor to prevent the brazing alloy from wetting the fixtures. Yet the parts to be brazed must remain wettable to insure free flow, even after two or three trips through the furnace.

MONEL* Alloy 404 is a nickel-copper composition in which residual elements are held to low levels, to provide a clean, wettable surface to the brazing alloy even after continued exposure to wet hydrogen. And it retains the basic properties of MONEL Alloy 400 . . . low vapor pressure, strength, ductility, corrosion resistance.

Brazed three times in wet hydrogen — The test assembly above illustrates the excellent brazing properties of MONEL Alloy 404. It was built up in three slow passes through the furnace using hydrogen with a dew point of +70°F. On the third pass, the brazing alloy still flowed the full length of the joint.

The low magnetic permeability of MONEL Alloy 404 is not significantly affected by processing and fabrication, making it particularly suited for electronic parts. In any commercial form, size, or temper the permeability will not exceed a value of 1.1 at 27°F when measured using a field strength of 0.5 oersted. A nominal or typical value is about 1.01 at 27°F. (1.004 at room temperature using a field strength of 200 oersteds).

The thermal expansion of MONEL Alloy 404 is about the same as that of MONEL Alloy 400, and is sufficiently close to that of many other alloys frequently used in the construction of radio tubes to permit the firing of composite metal structures with a negligible amount of distortion. MONEL Alloy 404 also retains much of its strength at "out-gassing" temperatures.

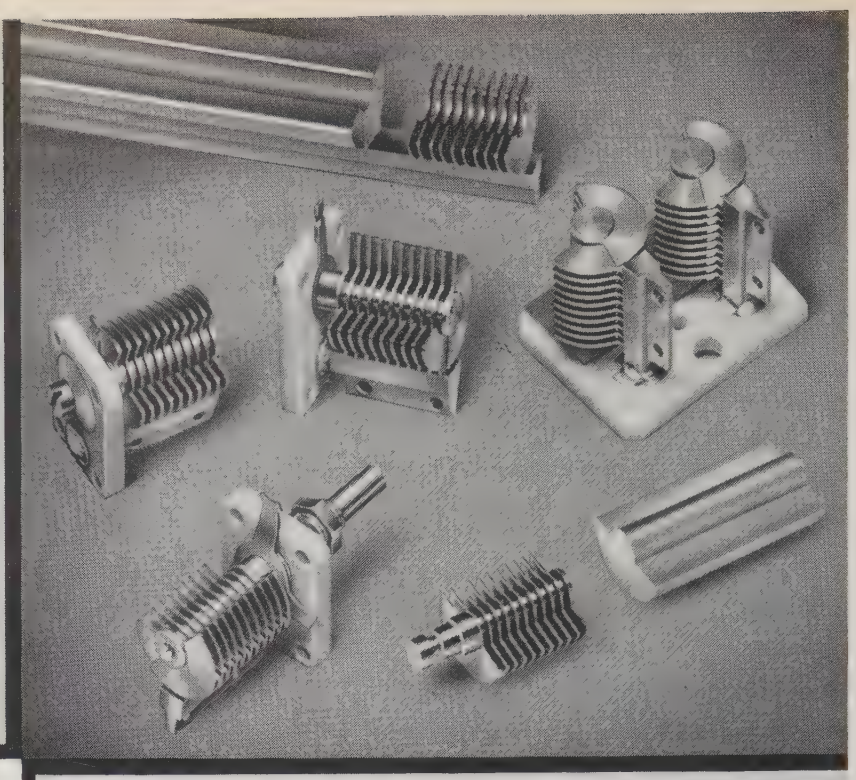
For more information about this new alloy—write for our bulletin—Basic Data — MONEL Alloy 404. *Registered trademark

HUNTINGTON ALLOY PRODUCTS DIVISION
The International Nickel Company, Inc.
Huntington 17 West Virginia

INCO

MONEL® Nickel-Copper Alloys

Designed for M Application



NO. 25000 SERIES HIGH Q MINIATURE VARIABLE AIR CAPACITORS

High Q at high frequencies. Rotors and Stators precision machined from extruded solid brass. Terminal integral part of Stator block. Shaft an integral part of Rotor. Screw-driver slot adjustment or ⅛" extended shaft for knob. Only 6 parts to a complete Capacitor. Stock sizes 15, 25 & 35 mmfd. single or dual units.

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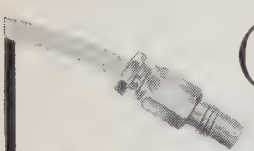
MASSACHUSETTS

RELIABILITY DELIVERED



NEW

SUBMINIATURE COAXIAL R F CONNECTORS

SMALLEST, LIGHTEST, MATCHED IMPEDANCE SUBMINIATURE CONNECTOR AVAILABLE

MICON, new as a company, old in experience, makes available the industry's most extensive line of uniquely designed bulkhead, chassis, line and printed wiring board connectors of the 50 ohm screw-on type.

We, at MICON, have prepared an evaluation kit which is available on request.

MICON ELECTRONICS, INC.

ROOSEVELT FIELD,

GARDEN CITY, L. I., NEW YORK

a wholly owned subsidiary of Metalcraft, Inc.

Professional Group Meetings

(Continued from page 90A)

Los Angeles—June 12

Commentary and Films by Charles Eames, joint meeting with STWP.

Northern New Jersey—June 8

"This English Language of Ours," J. Chapalino, Philco Corp., Philadelphia, Pa.

Information Theory

Washington, D. C.—April 27

"Least Square Synthesis of Radar Ambiguity Functions," S. M. Sussman, Melpar, Inc., Falls Church, Va.

Washington, D. C.—March 30

"Laboratory Error Detection for a Binary Communication System," A. E. Maser, Westinghouse Electric Corp., Baltimore, Md.

Microwave Theory and Techniques

Chicago—May 12

"Electron Beam Parametric Amplifier —Its Present and Future," Dr. R. Adler, Zenith Radio Corp., Chicago, Ill.

Los Angeles—March 22

"Applications of the Solid State to Microwaves," Dr. L. Hogan, Motorola, Inc., Phoenix, Ariz.; J. Cacheris, Motorola, Inc., Phoenix, Ariz.

Military Electronics

Northwest Florida—April 27

"A New Approach to Space Navigation," Maj. F. Setes, Eglin AFB, Fla.

Ottawa—April 27

"Possible Application of Optical Maser to Military Systems," Dr. H. Rothenberg, General Electric Co., Syracuse, N. Y.

Space Electronics and Telemetry

Central Florida—May 11

"Improved Demodulation and Predetection Techniques," Dr. L. R. Rauch, University of Michigan, Ann Arbor, Mich.

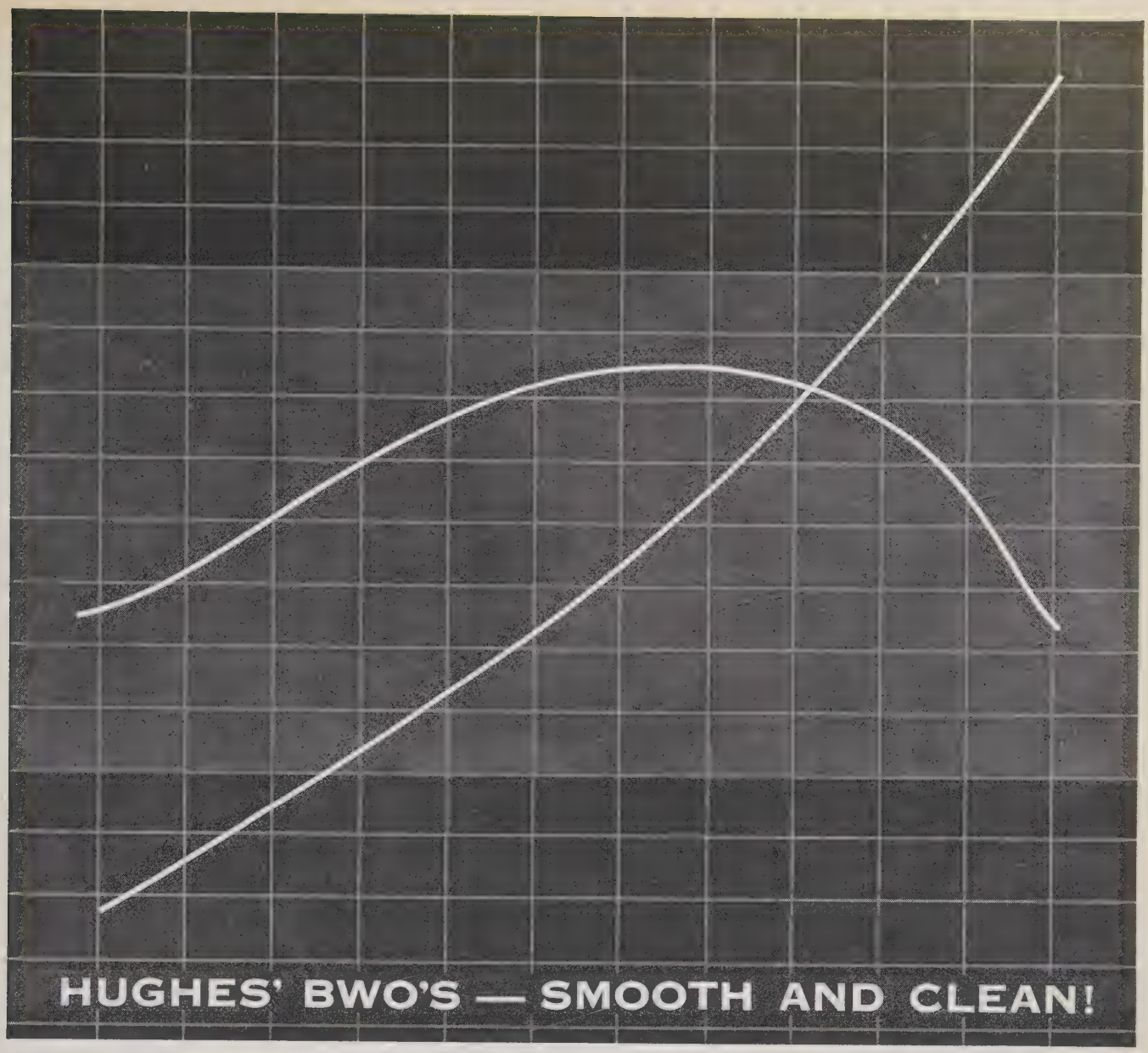
"Review of JETS Program and Future Plans," K. Mikalsen, Melbourne High School, Melbourne, Fla.

San Diego—June 28

"Micromodularization," W. B. Harris, RCA, Camden, N. J.

"Micromodularized Autopilot," I. A. Sofen, RCA, Los Angeles, Calif.

(Continued on page 94A)



Hughes' BWO's are smooth and clean!

Smooth—because fine grain power output variations are less than 1 db. Extremely smooth tuning curves make AFC easy. ■ Clean—because signal outputs with spectrum widths as narrow as .05 mc are typical, with commercially available power supplies. In phase-locked loops, signals of width well under 1 kc have been obtained. Noise 30 mc off carrier in two 1 mc bands was -96 dbm in a typical case. ■ Also—Hughes' tubes feature non-intercepting grids, low pushing and pulling factors, stability in phase-locked operation, and very high signal-to-noise ratios. ■ Pictured below are four popular models, available for immediate delivery, which cover the frequency range from 7 to 20 kMc. Further information on these or any special requirements for BWO's may be obtained from Hughes MTD.

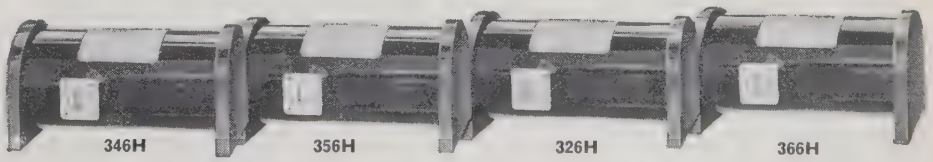
NORTHEASTERN 4 Federal Street, Woburn, Mass. WElls 3-4824 **EASTERN** 2000 "K" Street, N.W., Washington 6, D. C. FEderal 7-6760; 13 Lloyd Avenue, West Long Branch, N. J. CApital 2-1111 **WESTERN** 11105 South La Cienega Blvd., Los Angeles 45, Calif. SPring 6-1515

CREATING A NEW WORLD WITH ELECTRONICS

HUGHES

HUGHES AIRCRAFT COMPANY

MICROWAVE TUBE DIVISION



346H	356H	326H	366H
7.0—12.4 kMc	10.0—15.0 kMc	12.4—18.0 kMc	15.0—20.0 kMc
30 mw min P_o	10 mw min P_o	10—65 mw P_o	10 mw min P_o

Engineers who know
-SPECIFY
Q-max*
A-27
SUPERFINE
LOW-LOSS RF LACQUER

• Q-max, an extremely low loss dielectric impregnating and coating composition, is formulated specifically for application to VHF and UHF components. It penetrates deeply, seals out moisture, provides a surface finish, imparts rigidity and promotes stability of the electrical constants of high frequency circuits. Its effect upon the "Q" of RF windings is practically negligible.

• Q-max applies easily by dipping or brushing, dries quickly, adheres well; meets most temperature requirements. Q-max is industry's standard RF lacquer. Engineers who know specify Q-max! Write for new catalog.

*Registered Trademark

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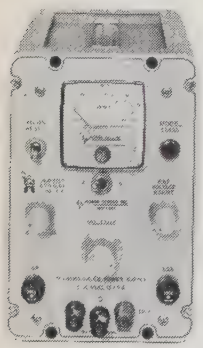
Telephone: HOpkins 2-3636



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MODEL
5015A

SEMICONDUCTORIZED
POWER SUPPLY
0-50 V.D.C. • 0-1.5 AMP

A Silicon Small-Signal Transistor Amplifier with Planar Diffused Junction Input Stage

High stability solid state DC power supply with adjustable current limiting, .05% regulation, 500 μv ripple, .01 Ω source impedance, 50 μs response time, .01 volts drift/24 hours, 55-440~ input.

$234.50
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IMMEDIATE DELIVERY
Write For Complete Catalog

Power Designs inc.
1700 SHAMES DRIVE
WESTBURY, L. I., N. Y.
EDgewood 3-6200 Area Code 516

Professional Group Meetings

(Continued from page 92A)

San Diego—April 12

"Synchronization Methods of Non-return-to-Zero Pulse Trains," Dr. M. H. Nichols, San Diego State College and Consultant.

San Diego—February 22

"Electrostatic Propulsion," Dr. D. G. Langmuir, STL Research Labs., Los Angeles, Calif.

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

Urban and Winget Appointed by Del

Michael R. Urban has been named sales manager at **Del Electronics Corp.** of Mount Vernon, according to Joseph Del-

cau, president. Urban was formerly military sales manager with Farrand Optical Co., Inc. He will be responsible for sales and advertising of Del engineered power supplies and other electronic equipment, as well as their new product line known as the Inductosyn®. This is a precision transducer having wide application for military use and automation control.

At the same time, Delcau announced the appointment of James L. Winget as manager of the newly-formed Inductosyn Department.

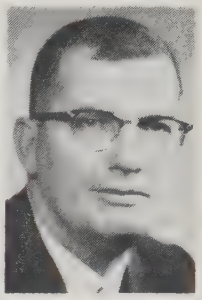
Winget has a masters degree in physics and electrical engineering from Oklahoma State University and 3 years of graduate work at California Institute of Technology. He served as a special research associate in infrared at Harvard University during the war. He was a consultant on problems of high vacuum deposition to Liberty Mirror Division, Libby-Owens-Ford Glass Co.

Winget was chief electronics engineer at Farrand Optical Co., where he was project engineer on the development of the Inductosyn and associated control equipment for machine tools, missiles and navigation systems.

For the last 18 months Winget has been senior staff engineer at Perkin-Elmer Corp.

Mullenbach VP of Electric Machinery

Robert J. Mullenbach has been appointed general manager of the **Control Div., Electric Machinery Mfg. Co.,** Minneapolis 13, Minn.



This is a new position resulting from the moving of the Control Department of E-M into larger facilities and its reorganization as a division.

Mullenbach will be responsible for all activities of the new division, including engineering, manufacturing and sales.

The division is responsible for the manufacturing of controls for motors and generators, static power converters and industrial control systems.

Prior to his appointment Mullenbach was a consultant and the General Manager of the Mullenbach Division in Los Angeles. E-M acquired the Mullenbach Electrical Manufacturing Company from Mullenbach and his brother Charles in 1954.

Trimmer Potentiometer

A new subminiature trimmer potentiometer, measuring $\frac{7}{16}'' \times \frac{1}{2}''$, for use in communications, computers, instrumentation, electro-medical equipment and other printed circuit applications, has been developed by **CTS Corp.,** Elkhart, Ind.

CTS Series 220 requires only $\frac{7}{16}'' \times \frac{5}{16}''$ printed circuit board area and is self-supporting, extending only $\frac{1}{2}''$ above the board surface. These units plug directly into

(Continued on page 96A)

Kodak reports on:

useful pictures this big ↓ ↑ a naive dream . . . f/1 lenses for 4.25μ to 10μ . . . Ektaline, sweet Ektaline . . .

Multum in parvo

Smaller than this. Oh, much smaller.

Smaller even than this.

→ · ← In fact, in this small an area it is becoming possible today to construct useful apparatus of considerable complexity.

It is done by photography. The apparatus is drawn to a convenient scale. The drawing is photographically reproduced in the requisite microscopic size. The tiny photograph of the drawing of the apparatus is converted by various techniques to the apparatus itself. This is the principle.

The details take a bit of doing. So bright are the prospects of reward for working them out that many are applying themselves to the effort with much vigor and encouraging success. An understanding of a few optical and photographic fundamentals of microphotography helps. For a brief summary of these fundamentals write Eastman Kodak Company, Special Sensitized Products Division, Rochester 4, N. Y. No specific directions for getting rich or famous are supplied.

A sharp eye for infrared

The decision to announce at this time the availability of *f*/1 *Irtran-2 Aspheric Lenses* has been reached in struggle against deeply rooted inhibitions. In the photographic film and paper trade we are habituated to a longer silence before the first blast of the trumpets. Infrared technology hates to wait, however, and puts a heavy load on the phone company's long lines.

These Irtran-2 lenses transmit usefully from 2μ to 14μ. Two focal lengths, 2-inch and 3-inch, are offered off the shelf. At *f*/1, we seem to have done well at providing high collecting-power for energy without undue sacrifice of sharpness. Sharpness was the goal. For both lenses, the minimum circle of confusion *computes* at less than .001″ for any wavelength from 4.25μ to 10μ. The italics mark where we hurt.

Much care and a valuable ingenuity have been exercised in impressing our tenth-degree equation upon the concave side of these meniscus lenses, in grinding and polishing the spherical convex side, in placing the center of the spherical curvature on the axis of the asphere, in maintaining the center thicknesses at the 9.1 mm and 10.4 mm values respectively that the calculations assume, in the optical homogeneity of the Irtran-2 material. More than this we cannot claim. To the extent that the care and ingenuity have succeeded in making the calculations represent the actuality, the circle of confusion is less than .001″. The customer's willingness to take a chance that we have hit it will, in good sense, depend on how badly his project needs a 2μ-14μ infrared image of high definition and high aperture.

To demonstrate experimentally at those wavelengths that the circle of confusion is indeed that small is a task which we have simply been too busy to complete up to the time these words were written.

In the lead sulfide region, the sharpness does not compute to be as good as farther out in the infrared. Yet we have customers who use the lenses there and are happy with confusion-circle minima as large as .008″.

In comparison with reflective optics hitherto used, Irtran-2 aspheres offer compactness and a wider field that doesn't even show any appreciable deterioration as far as 2° off axis. You do give up the perfect achromatism of reflective optics. In the 2-inch lens the minimum circle of confusion for 10μ radiation is located 2 mm beyond the minimum circle of confusion for 4.25μ radiation; in the 3-inch lens the separation is 3 mm.

We have said enough to establish our frankness and to indicate whether there is any need for you to burden the long lines to Rochester, N. Y., LOcust 2-6000, Extension 5166, which is one way to reach Eastman Kodak Company, Special Products Division. Bear in mind that Irtran-2 material has a hardness of 354 Knoop, is not at all brittle, withstands thermal shock and the solvent action of water, and can get very hot without losing transparency or giving off toxic fumes.

THIS paper

"*My husband sells oscillograph paper. Competition is fierce. He comes home beat every night.*"

Few overhearing her would know what the poor soul is talking about, yet she speaks the truth. With research and development activity now constituting such a respectable fraction of the Gross National Product, oscillographs probably outnumber pickle barrels in this country at the present time. Oscillographers are correspondingly numerous. Methods that one sect of oscillographers prefers above all else another sect can't see for dirt. One sect prefers automatic oscillogram processors. Paper manufacturers like us find their favor worth competing for. Therefore we announce a new advance in media for their use.

An advance in the old art of papermaking came first. Then new emulsions were devised to work properly with the new base. Then proper processing chemicals were devised for the new emulsions. Then the combination was extensively proved out under practical conditions of use by parties interested only in end results and hardly at all in the how and why. They found that

1. THIS paper dries thoroughly at high processor speeds *without creases.* ***180 in./min. is not too fast.***
2. THIS paper gives trace lines that stand out as black as the ace of spades. ***Background is nice and clean.***
3. THIS paper isn't fussy about how long it sits around before use. ***O.K. to keep plenty on hand.***
4. THIS paper is rugged. ***No cracking, no crumbling.***
5. THIS paper holds its dimensions. ***Justifies careful measurement.***

"*THIS*" *won't do for a trademark.* (*The code name for the field trials was "Kind 1534.") Let's call it* Kodak Ektaline Paper. *It comes in the two usual speeds for oscillographs,* Kodak Ektaline 16 Paper *and* Kodak Ektaline 18 Paper. Kodak Ektaline Chemicals *come as liquids. The stabilization principle used in the automatic oscillogram processors came from Kodak, too. An inquiry to Eastman Kodak Company, Photorecording Methods Division, Rochester 4, N. Y., puts everything in place right up to the moment.*

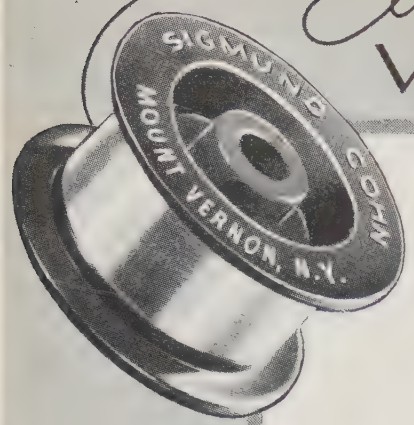
How to stick together

The naive dream of a stickum to stick metal to metal without heat, waiting, clamping, or shrinkage came true in *Eastman 910* adhesive. It is the costliest on the market. It can bond almost anything—not just metal—to almost anything else. Hundreds of sample-buyers have solved serious assembly problems with the stuff. Techniques have evolved. They have to be seen to be believed. To show some of them, we have made a 15-minute sound movie for showing to professional and industrial groups. It demonstrates how-not-to's along with the how-to's.

To borrow the film for a showing, write Eastman Chemical Products Inc., 260 Madison Avenue, New York 16, N. Y. (*Subsidiary of Eastman Kodak Company*).

Kodak
TRADE MARK

This is another advertisement where Eastman Kodak Company probes at random for mutual interests and occasionally a little revenue from those whose work has something to do with science



Electroplated WIRES

Often, in the manufacture of formed wire parts it is more economical to form these parts from wire already electroplated than it is to plate the formed pieces after manufacture... Our facilities for continuous electroplating are regarded as the largest in the field. All our plating is consistently uniform and well-bonded to the base wire... Consult us about any specific electroplating problem.

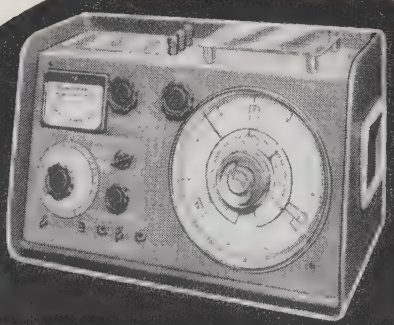
Write for latest brochure.

SINCE 1901

SIGMUND COHN MFG. CO., INC.

121 SOUTH COLUMBUS AVENUE, MOUNT VERNON, N.Y.

SIGMUND COHN CORP. OF CALIFORNIA • 151C N. Maple St., Burbank, Cal.



Model 1313

$590

1/4% ACCURACY WANTED?

This new Universal Bridge adds to the wide variety from which an engineer must choose. But Model 1313 has both ¼% Accuracy and Direct Readout; combines high discrimination with exceptional ease of use. Detector AGC, variable frequency operation and functional styling are all plus features.

L: . . . . . 1μH to 110H, 7 decades
C: . . . . . 1μμF to 110μF, 7 decades
R: . . . . . .01Ω to 110MΩ, 8 Decades
Accuracy: . . . ¼%
Discrimination: 5000 div'ns/Decade
Frequency: 1Kc, 10Kc. 100 cps to 20Kc with ext. osc.
Readout:. . . . Direct—no multiplying factors

MAKE NO MISTAKE
MEASURE WITH:

MARCONI INSTRUMENTS

111 CEDAR LANE • ENGLEWOOD, NEW JERSEY

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 94A)

board and can be used in equipment where multiple boards are stacked on $\frac{5}{8}$″ centers. Resistance range is 250 ohms thru 2.5 megohms (linear taper, other tapers available). Wattage ratings are $\frac{1}{8}$ watt @ 55°C derated linearly to no load @ 85°C with control mounted on the board. Voltage rating, shaft to terminals, is 750 vac for one minute high pot test, 500 vdc operating maximum; 350 vdc across end terminals, load not to exceed wattage rating. Delivery is two weeks for sample quantities, four weeks for production quantities. Price is approx. 90¢ each in production quantities.

Bailey New VP For Marconi

Marconi Instruments, 111 Cedar Lane, Englewood, N. J., has become a division of the newly formed **English Electric Corp.**

R. J. (Bill) Bailey has been appointed vice president and will continue as manager of the **Marconi Instruments Division.**

Bill Bailey joined Marconi Instruments in 1938 as a junior Sales Engineer and was Manager of their London office before coming to the USA in 1954.

Sixty Message Capacity Readout For Annunciator

Sixty messages can be displayed individually, or in any combination, by the new Series 90H display for annunciators which was introduced by **Industrial Electronic Engineers, Inc.,** 5528 Vineland Ave., N. Hollywood, Calif., makers of a line of readout devices.

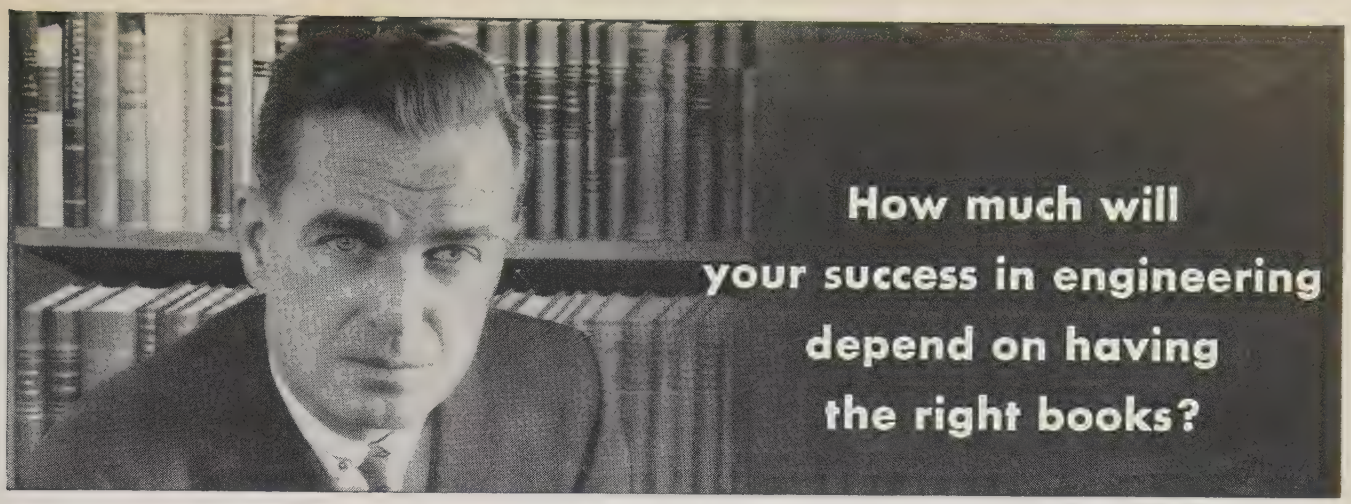
Red or green color backgrounds are available in units with a fifty message capacity.

Through its rear-projection design all data (for example, "Temperature, Stop-Go, Test, Numbers, or any characters of varying length, width, height, or color) appear on one plane on the front surface of the unit.

Data is of uniform intensity and readable from any viewing angle. Compact over-all size of the 90H is 5½″ wide, 12″ long, and 16½″ high.

For more complete information, write direct to the manufacturer.

(Continued on page 120A)



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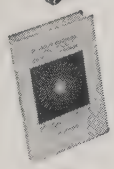
Start your membership with any of these selections:

Transistor Electronics by D. DeWitt and A. L. Rossoff. Full coverage from basic fundamentals to detailed mechanisms.

Publisher's Price, $9.00
Club Price, $7.65



Electronic Digital Computers by C. Smith. What they are, how they function, and their underlying principles.
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Control Engineers' Handbook by J. C. Truxal. A wealth of practical help on automatic feedback control systems.

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Mathematics for Electronics with Applications by H. M. Nodelman and F W. Smith, Jr. Methods for solving practical problems.

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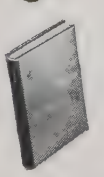
Epoxy Resins by H. Lee and K. Neville. Fully treats their modern applications and technology.

Publisher's Price, $8.00
Club Price, $6.80

How to Become a Professional Engineer by J. D. Constance. Authoritative guidance applicable in all states.

Publisher's Price, $5.50
Club Price, $4.70



Management for Engineers by R C. Heimer. Practical know-how for the engineer-manager, covering every management area.

Publisher's Price, $7.50
Club Price, $6.40

Your engineering career owes a great deal to books. Why not take advantage of this convenient, economical way to have the best professional books available when you need them? THE ELECTRONICS AND CONTROL ENGINEERS' BOOK CLUB brings you the essential technical literature in your field. It also helps you overcome today's high cost of building a professional library *by saving you an average of 15% from publisher's prices.*

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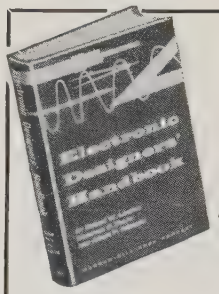
Send No Money Now. Just check the book you want as your first selection on the coupon below. With it you will be sent *Electronic Designers' Handbook* for only one dollar. Take advantage of this offer and receive two books for less than the regular price of one.



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for only **$1.00**
with membership in
The Electronics and Control Engineers' Book Club

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By R. W. Landee, D. C. Davis, and A. P. Albrecht

Provides fundamentals and data to help you in the design of all types of electronic equipment. A large number of the circuits used in many different applications are covered, together with theoretical and technical discussions and explanations, design examples to show application of theory, and graphical and tabular data needed in day-to-day design work.

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Please enroll me as a member of The Electronics and Control Engineers' Book Club. You will bill me for my first selection indicated at right at the special club price and $1 for **Electronic Designers' Handbook**, plus few cents delivery costs. (The Club assumes this charge on prepaid orders.) Forthcoming selections will be described in advance and I may decline any book. I need take only 3 selections or alternates in 12 months of membership. **(This offer good in U.S. only.)**

PLEASE PRINT

Name .....

Address .....

City .....Zone .....State .....

Company .....

Send as my first selection:

- ☐ Transistor Electronics, $7.65
- ☐ Electronic Digital Computers, $10.25
- ☐ Control Engineers' Handbook $15.75
- ☐ Mathematics for Electronics with Applications, $5.95
- ☐ Epoxy Resins, $6.80
- ☐ How to Become a Professional Engineer, $4.70
- ☐ Management for Engineers, $6.40

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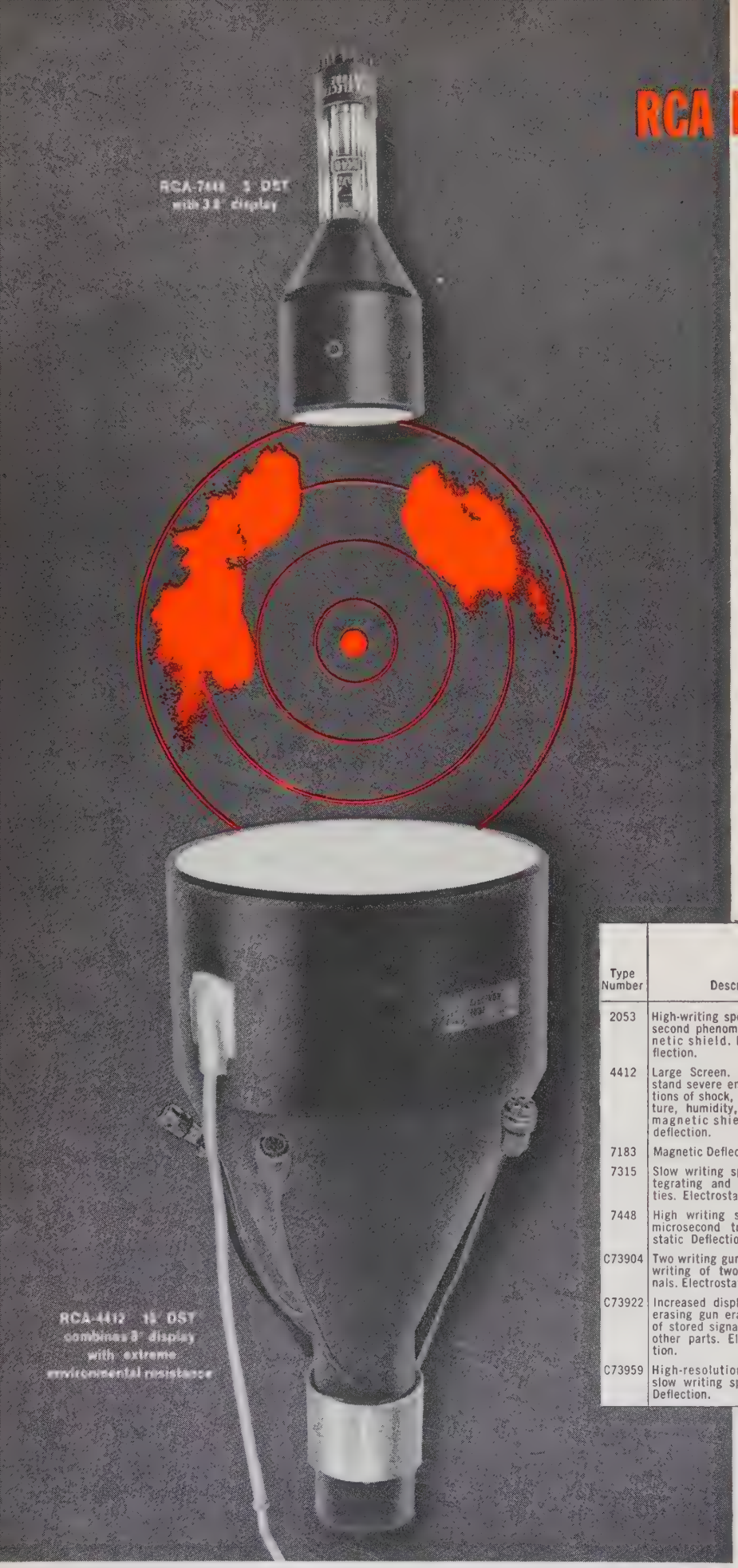
Write your own specifications for

RCA Display Storage Tubes



Fill in your requirements in a Display Storage Tube—then leave the rest to RCA. With RCA DST's, you have widest possible latitude in your choice of

Characteristic	Your requirement
Writing speed	_____
Size of display	_____
Degree of environmental resistance	_____
Degree of resolution	_____
Number of writing guns	_____
Type of deflection	_____
Collimation simplicity	_____
Selective erasure	_____
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7183	Magnetic Deflection.	5.19	4	–	12000	50
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7448	High writing speed to "freeze" microsecond transients. Electrostatic Deflection.	5.31	3.8	300,000	2500	50
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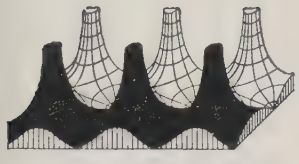
Poles and Zeros



E&A-ECPD. These symbols stand for the Education and Accreditation Committee of the Engineers Council for Professional Development. In July, 1961, Poles and Zeros announced IRE's membership in ECPD and discussed briefly the ECPD committee structure. The Education and Accreditation Committee is of such significance that it seemed appropriate to single it out for detailed discussion on this page. In order to provide the most accurate and precise information about the committee, the Editor invited the committee's chairman to prepare a statement descriptive of its activities for Poles and Zeros.

Newman A. Hall, Chairman of the Department of Mechanical Engineering at Yale University, is the Chairman of the ECPD E&A Committee, and he has graciously accepted the Editor's invitation to prepare for IRE readers the story of the responsibilities and activities of his committee. For the material to follow, IRE and the Editor express their thanks to Professor Hall.

"The addition of the IRE as one of the constituent members of ECDP brings with it the obligation and opportunity to participate in one of its most significant activities, the Education and Accreditation Committee. In the total responsibility of ECPD in the professional development of the engineer, this committee plays the central role of maintaining contact with the institutions which provide the formal education of the engineer. It has the responsibility of speaking for the profession in matters of educational standards and of developing and administering criteria so that a reliable and meaningful program of accreditation of engineering schools can be carried forward.

"The establishment of accreditation criteria and an associated inspection program for engineering schools was one of the earliest formal programs of ECPD. The basic criteria were agreed upon in 1933 and, during the period 1936 to 1938, the initial program of accreditation was completed. During the war, the inspections of curricula were largely suspended and, on resumption afterwards, it was observed that a much more adequate base for conducting accreditation was needed. While the earlier Hammond Reports of 1940 and 1944 on the aims of engineering curricula led the way for changes which were taking place, a more extensive review was in order. Consequently, following a 1951 ECPD recommendation, ASEE instituted a study which resulted in the 1955 report on The Evaluation of Engineering Education. With this available, ECPD adopted in 1955 a set of additional criteria for accreditation which, together with the original bases, have since guided the recommendations of the Education and Accreditation Committee. These standards which the committee administers are designed to accommodate and encourage flexibility and creative development in engineering education, yet they are sufficiently incisive to insure that minimum requirements are significant.

"The membership of the Education and Accreditation Committee is representative of all the constituent societies of ECPD, including IRE. Great care is taken not only to insure that the committee is broadly familiar with various technical fields of engineering, but more significantly, that the individual members may be in a position, on the basis of their own experience in engineering education and good judgment, to counsel educational institutions wisely. Members in all cases are selected from individuals who have had experience as inspectors on accreditation teams, and it is the responsibility of each constituent society to insure that an adequate corps of representative, well-informed inspectors is available to the committee.

"Each accreditation inspection is organized by a member of the E&A Committee who serves as chairman. He is responsible for making arrangements with the institution, selecting inspectors from approved lists, supervising the inspection, including various conferences with administrative officials, and assembling and transmitting reports and recommendations to the E&A Committee for action. The final accreditation decision is made by ECPD itself, giving due consideration to the recommendations of the E&A Committee.

"A responsibility is accepted, not only to insure that minimum standards are maintained by the engineering schools, but increasingly in recent years there has been an effort to provide constructive assistance to the institutions in their constant efforts to improve the quality of their educational program. With the introduction of new principles and concepts into engineering practice, and with new proven methods of training engineers always becoming available, this service can be one of the most important provided by the committee.

"In order for such broad services to be possible, a very close relation must be maintained between the constituent societies and the committees. This is accomplished very effectively by the informal contacts provided by the membership of the committee in the several societies. Generally, the members are selected from those already very active in education and society affairs.

"With the addition of new organizations to ECPD, the scope of responsibility is broadened. The increased diversity in technical areas and operating levels in the engineering team effort constantly enlarges the area of concern. The goal of the committee is to effectively encompass all of these demands and to provide the maximum service to the profession."

—F. H., Jr.

Daniel E. Noble

Director, 1961

Daniel E. Noble (A'25–SM'44–F'47) was born in Naugatuck, Conn., on October 4, 1901. He graduated from Connecticut State College, Storrs, and supplemented his studies with additional work at Harvard Summer School and at M.I.T., both at Cambridge, Mass. An honorary D.Sc. degree was conferred upon him in 1957 by Arizona State College at Tempe. He taught mathematics and electrical engineering subjects for seventeen years before he joined Motorola, Inc. in 1940, as Director of Research. Before leaving the University of Connecticut, he pioneered FM radio broadcast and mobile radio development work and, as consultant to the Connecticut State Police, he was responsible for the design and development of the first statewide two-way mobile radio communications state police system and the first FM system. With Motorola, he advanced over the years to General Manager of the Communications and Electronics Division, to Vice President and Director of the corporation, and, finally, to his present assignment as Director and Executive Vice President in charge of the three Motorola Divisions dealing with communications, semiconductor and military electronics products. While Manager of the Communications Division of Motorola, he directed the development of equipment improvements which paced the universal changeover of land mobile radio systems from AM to FM. He particularly emphasized engineering effort directed to the solution of the problems related to the need for substantial increases in the efficiency of spectrum utilization.

With the announcement of the invention of the junction transistor, his interest turned to the semiconductor and solid state electronics fields as the most promising areas for corporate expansion. He initiated programs of transistor circuitry research and transistor applications in the Communications and Military Electronics Divisions, and established the corporate semiconductor research organization, which was later expanded into the Motorola Semiconductor Products Division. He also organized the Solid State Electronics Department which extended the research into the areas of specialized solid state devices, integrated circuits, and informational processing and controls.

Mr. Noble has served either as Chairman or as a member of many industry and professional technical and advisory committees, including the Radio Technical Planning Board (Chairman of Panel 13), and the National Color Television Systems Committee. He holds three patents on electronic devices, and he has published technical articles and papers dealing with the general problems relating to the development of effective and reliable equipment. He is a member of the Board of Directors of Arizona State University Foundation, and Chairman of the Foundation Advisory Committee on Engineering and Science Education. He is also a member of the Army Scientific Advisory Panel.

Scanning the Issue

Time-Varying Networks, I (Zadeh, p. 1488)—It has been the practice of the PROCEEDINGS to publish from time to time review and tutorial articles by leading authorities on subjects in which important progress has recently been made. Earlier this year the Editorial Board of the IRE set into motion a program for stepping up the frequency with which these articles appear. This paper was suggested for this program by the Professional Group on Circuit Theory and represents the first fruit of the Editorial Board's efforts. It deals with a subject about which very little was written prior to 1950. The past decade, however, has seen a host of technological developments that have given rise to problems involving the analysis and synthesis of time-varying systems. By way of example one could mention the advent of Wiener's theory, missile guidance techniques, parametric amplification, scatter communication, detection of fluctuating targets, communication via satellites, and many more. Presented here is Part I of an excellent two-part survey of some of the significant and lesser known contributions to time-varying-network theory that have been made in the U. S. and abroad during the past ten years.

Compatible Single Sideband (Kahn, p. 1503)—Due to the ever-increasing congestion of the radio frequency spectrum there has been widespread interest in recent years in single-sideband communication techniques. Indeed, interest in SSB reached such a level that the PROCEEDINGS devoted a special issue to the subject in December, 1956. Meanwhile, in August of that year, a new SSB system was quietly being installed at the U. S. Voice of America station in Munich, Germany. The new system was unique in that it did not require special SSB receivers. The broadcasts could be satisfactorily received with conventional AM broadcast receivers. In the five years that have intervened, the Compatible Single Sideband (CSSB) system has been the center of considerable controversy—controversy that has reached the pages of the PROCEEDINGS on two occasions. Questions were raised as to whether a CSSB wave with a distortion-free envelope could reproduce speech and music with adequate fidelity without generating an appreciable amount of undesired sideband radiation. These discussions arose largely because the full details of the system had never been disclosed for proprietary reasons. In the meantime, the CSSB system was installed and tested at five AM broadcast stations in the U. S. as well as at various government and aeronautical installations in a number of countries. The author has now set down a detailed description, explanation, and experimental evaluation of the CSSB system. This paper, then, represents an eagerly awaited first public disclosure of an important new single-sideband system that has been the subject of much interest and discussion among a wide circle of engineers in the broadcast and communications field.

Magnetic Circuits Employing Ceramic Magnets (Helmer, p. 1528)—Ceramic magnets are characterized by the fact that they have a fixed magnetization and a permeability equal approximately to that of free space. These properties lead to an interesting simplification of the field equations, a simplification which makes it readily possible for a device designer to design magnetic circuits to his own specifications rather than relying on a magnet manufacturer for the design. The author analyzes several types of magnetic circuits, including novel ceramic-steel and ceramic-iron configurations, and develops a method of optimizing their design. The paper could well be called semitutorial since, in addition to formulating design criteria, it brings a not-well-understood subject within ready grasp of applications people.

Electron-Trigonometry—A New Tool for Electron-Optical Design (Schlesinger, p. 1538)—The designing of an electron-optical system is basically a problem of predicting the size and position of electron images. Unfortunately, in most situations of practical interest it is not a problem that can be readily described or easily and directly solved mathematically. The designer, therefore, usually must resort to indirect analog techniques or tedious graphical methods of tracing principal trajectories, or analyze the problem with a computer in a way that, while it gives a specific answer, provides no insight into design relationships. In this paper the author develops a new method in which axial potential distributions of an electron-optical system are approximated by segments that are either linear or parabolic. This approach reduces the problem to trigonometric relationships that are not only simpler to handle but more direct than present methods of analysis. Moreover, the method lends itself well to programming for a digital computer. The wide application of electron optics today makes this a noteworthy contribution to the art.

A New FM-AM Method of Compression, Expansion, and Multiplication (Aiken and Süsskind, p. 1550)—This brief paper describes a novel method of combining two signals in such a way that the dynamic range of the first signal is expanded or compressed by the second. The technique has particular application to controlling the average levels of audio, radio or television signals. However, it may also find other interesting uses where modulation with separated components is involved, for example, as a multiplier in analog computers.

IRE Recommended Practices on Audio and Electroacoustics: Loudspeaker Measurements (p. 1553)—This document, drawn up by the Audio and Electroacoustics Committee of the IRE, defines terms associated with loudspeakers and their testing, recommends various methods of testing, and indicates preferred methods of presenting information regarding their characteristics. Included are methods of measuring reference pressure response, rating efficiency, directional properties, nonlinear distortion, and power capacity rating.

Coincidence Techniques for Radar Receivers Employing a Double-Threshold Method of Detection (Endresen and Hedemark, p. 1561)—The authors propose that the well-known double threshold radar detection technique requiring m or more detections in n consecutive trials be replaced by a video coincidence technique in which m or more consecutive *pairs* (or triples) of detections are required in n consecutive trials before saying that a target is present. The significance of requiring that the detections be clustered in groups of two or three is that this clustering is exactly what one expects when a true target appears and, on the other hand, is not at all characteristic of independent and random noise pulses which give rise to false alarms. Mathematical formulas are developed to determine how much is gained in terms of reducing false alarm probabilities and what price is paid in terms of reducing the detection range. The answers to these questions and the proposed coincidence technique itself will be of substantial interest to those concerned with suppressing pulse-type interference in radar systems.

Time-Varying Networks, I*

L. A. ZADEH†, FELLOW, IRE

Summary—**This paper presents an analysis of some of the significant developments in time-varying network theory which have taken place during the past decade, with the emphasis placed on three topics: 1) characterization of time-varying networks, and in particular, transition from the impulsive response to the differential equation; 2) the problem of factorization, with emphasis on the contributions of Darlington, Batkov and Paul Levy; and 3) randomly-varying systems and, in particular, the question of stability of discrete-time systems of this type. The identification problem, the analysis of periodically-varying systems, the synthesis problem, and the filtering and prediction of nonstationary processes will be treated in Part II, to be published later.**

THIS PAPER is primarily concerned with developments in time-varying-network theory which have taken place during the past decade. A survey paper by Bennett [1] provides a brief discussion of the developments prior to 1950.

The past ten years have witnessed a rapid growth of interest in time-varying networks as well as a marked shift in emphasis brought about by the advent of such new areas of research as missile guidance, detection of fluctuating targets, propagation through randomly-varying media, communication via satellites, parametric amplification, etc. Furthermore, the availability of machine computers has made, and is making, a profound impact on the whole process of constructing a model of a physical system and subjecting it to mathematical analysis, as well as on the reverse process of synthesizing a system from components having known characteristics. One consequence of the advent of machine computers is the growth in importance of the so-called identification problem, which, roughly speaking, is concerned with the determination of the characteristics of a given black box, based upon the observation of its external behavior. This stems from the fact that, with the aid of machine computers, the analysis of a system comprised of elements having known characteristics no longer presents essential difficulties. Thus the outstanding problem becomes that of identification, which in any case is a basic prerequisite to the analysis of a system by either machine computer or analytical means.

This paper is not purported to be an exhaustive survey. Rather, its limited aim is to present a connected and not too superficial account of some of the significant and lesser known contributions to time-varying-network theory made during the past ten years. Because of limitations on space, the paper is in two parts. Part I deals with the problem of transition from one mode of characterization to another, the so-called factorization problem, and with randomly-varying systems. The identification problem, the analysis of periodically-varying systems, the synthesis problem, the stability problem, the filtering and prediction of nonstationary processes, the resolution into elementary time functions and miscellaneous contributions will be treated in Part II, to be published later.

It should be noted, in the following exposition, that the amount of space devoted to a particular contribution is not necessarily a measure of its significance. As a general rule, the contributions which are well known or have appeared in widely available journals are referred to very briefly, while the results which appeared in unpublished reports or foreign publications are discussed in greater detail.

I. Characterization

It has long been known that the relation between the input and output of a time-varying system can be expressed in a variety of ways other than those based on the use of differential equations. Yet it was—and to some extent still is—standard practice to employ almost exclusively the differential-equation representation for the purpose of relating the input to the output, either directly or through the state variable.

It is largely during the past decade that a need for a wider variety of alternative representations became definitely established and their potentialities as well as limitations became more clearly understood. The reason for this development is twofold: first, it was found that many of the time-varying models of such systems as randomly-varying media, fluctuating targets, etc., cannot be characterized in terms of ordinary differential equations of finite order. Second, it became clear that in many of the problems of optimization, identification, detection, filtering, etc., it is much more convenient to use a representation tailored to the characteristics of a particular class of systems than to use an "all-purpose" representation involving a differential equation or a set of differential equations relating the input to the output.

In speaking of the modes of characterization other than those based on the use of differential equations, what we have in mind are the techniques centering on the resolution of input time functions into a family of "elementary signals" such as delta functions, exponentials, etc. If the responses of the system to such elementary signals can be determined, then the response to

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an arbitrary input can be determined by superposition. In this way, the problem of finding the response to a given input is reduced to a large number of simple problems, each involving the determination of the response of the system to a particular elementary signal.

Before proceeding to discuss some of the contributions to the characterization of time-varying systems, it would be helpful to summarize some of the well-known facts pertaining to the technique of resolving the input and output into elementary signals. (For a more detailed discussion, see Huggins [19], Laning and Battin [16], Gerardi [12], and Zadeh [2].)

Resolution into Delta Functions

In this case, the family of elementary signals has the form $\{\delta(t-\xi),\ -\infty<\xi<\infty,\ -\infty<t<\infty\}$ where $\delta(t-\xi)$ denotes a unit impulse occurring at time ξ, with ξ ranging over the interval $(-\infty,\ \infty)$. Assuming, for simplicity, that the system has a single input u and a single output y, and that it is initially at rest, the relation between the input and output can be written as

$$y(t) = \int_{-\infty}^{\infty} h(t, \xi)u(\xi)d\xi \tag{1}$$

where $h(t, \xi)$ is called the *impulsive* (or impulse) *response* of the system and is defined as the response of the system at rest to $\delta(t-\xi)$. If a system is *nonanticipative* (*i.e.*, if, for all t, $y(t)$ is independent of the values of the input for times greater than t), then $h(t, \xi) \equiv 0$ for $t<\xi$ and the upper limit in (1) can be replaced by t. (For a discussion of the relationship between the impulsive response and Green's function, see Miller [3] and Zadeh [4].)

Resolution into Exponential Functions

Here, the elementary signals are of the form $\{e^{j\omega t},\ -\infty<\omega<\infty,\ -\infty<t<\infty\}$ or more generally $\{e^{st},\ s\in C\}$ where C is a Bromwich-Wagner contour in the s plane. In terms of such signals, a system B is characterized by its frequency-response function $H(j\omega, t)$, which is defined as

$$H(j\omega, t) = \frac{\text{response of } B \text{ to } e^{j\omega t}}{e^{j\omega t}}. \tag{2}$$

The time-varying frequency-response function $H(j\omega, t)$ constitutes a natural generalization of $H(j\omega)$—the frequency-response function of a time-invariant system. If the system is initially at rest and the Fourier transform of the input u is denoted by $U(j\omega)$, the expression for the output at time t, $y(t)$, in terms of $H(j\omega, t)$ and $U(j\omega)$ is

$$y(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty} H(j\omega, t)U(j\omega)e^{j\omega t}d\omega. \tag{3}$$

General Relations for Resolution into Elementary Time Functions

If a family of time functions $\{k(t, \lambda)\}$, with λ taking values on some contour C in the λ plane and t ranging over a finite or infinite interval T, form a basis for the input and output spaces of a system B [in the sense that the $k(t, \lambda)$ and C are such that any signal in the input or the output space of B can be resolved into the $k(t, \lambda)$], then an input u can be expressed as

$$u(t) = \int_C k(t, \lambda)U(\lambda)d\lambda, \qquad t \in T, \tag{4}$$

where $U(\lambda)$ is an integral transform of u given formally by

$$U(\lambda) = \int_{-\infty}^{\infty} k^{-1}(\lambda, t)u(t)dt. \tag{5}$$

In this relation, the kernel $k^{-1}(\lambda, t)$ is inverse to $k(t, \lambda)$ in the sense that

$$\int_C k(t, \lambda)k^{-1}(\lambda, \xi)d\lambda = \delta(t-\xi), \qquad t, \xi \in T. \tag{6}$$

B is characterized by its responses to the $k(t, \lambda)$, $\lambda \in C$. Thus, if the response of B to $k(t, \lambda)$ is denoted by $K(t, \lambda)$, then the response of B to u (with B initially at rest) can be written as:

$$y(t) = \int_C K(t, \lambda)U(\lambda)d\lambda. \tag{7}$$

Eqs. (1) and (3) are special cases of this relation corresponding to $k(t, \lambda) = \delta(t-\lambda)$ and $k(t, \lambda) = e^{j\lambda t}/2\pi j$, respectively.

Resolution into Eigenfunctions

A function $k(t, \lambda)$, $t \in T$, is an eigenfunction of B if the response of B to $k(t, \lambda)$ (with B initially at rest) is of the form: constant (depending on λ) times $k(t, \lambda)$. If B has a set of eigenfunctions $\{k(t, \lambda), \lambda \in C\}$ which can be used as a basis for the input and output spaces of B, then the input-output relationship of B can be expressed in the simple form

$$Y(\lambda) = R(\lambda)U(\lambda) \tag{8}$$

where $R(\lambda)$ is the response function of B

$$R(\lambda) = \frac{\text{response of } B \text{ to } k(t, \lambda)}{k(t, \lambda)} \tag{9}$$

and $U(\lambda)$ and $Y(\lambda)$ are the integral transforms of u and y, respectively; *i.e.*,

$$U(\lambda) = \int_T k^{-1}(\lambda, t)u(t)dt \tag{10}$$

and

$$Y(\lambda) = \int_T k^{-1}(\lambda, t)y(t)dt. \tag{11}$$

It can be shown readily that if a system is time-invariant and homogeneous (in the sense that if u is any input in the input space of B, then for all real constants k, the response of B to ku is k times the response of B to u), then the functions $\{e^{j\omega t},\ -\infty<\omega<\infty,$

$-\infty < t < \infty\}$ constitute an eigenfunction set for B. (Note that B need not be linear.) It is in this sense that the exponential functions $\{e^{j\omega t}\}$ constitute a "natural" set of elementary time functions for linear time-invariant systems.

This concludes our brief review of the terminology and basic relations pertaining to the characterization of linear systems in terms of their responses to elementary time functions.

II. Transitions Between Different Modes of Characterization

In dealing with time-varying systems, one frequently encounters a situation where a system B is characterized in a given way W, and it is desired—for one reason or another—to characterize B in a particular way W^* which is different from W. For example, B might be initially characterized[1] by a differential equation of the form

$$a_n(t)\frac{d^n y}{at^n} + \cdots + a_0(t)y = b_m(t)\frac{d^m u}{at^m} + \cdots + b_0(t)u, \quad (12)$$

or more compactly

$$L(p, t)y = M(p, t)u, \qquad p = \frac{d}{dt}$$

where

$$L(p, t) = \sum_{i=1}^{n} a_i(t)p^i$$

$$M(p, t) = \sum_{j=1}^{m} b_j(t)p^j$$

and the $a_i(t)$ and $b_j(t)$ are given functions of time. The desired characterization of B is:

$$y(t) = \int_{-\infty}^{\infty} h(t, \xi)u(\xi)d\xi \quad (13)$$

where $h(t, \xi)$ is the impulsive response of B. Here, the problem is that of solving the differential equation

$$(a_n(t)p^n + \cdots + a_0(t))h(t, \xi) = (b_m(t)p^m + \cdots + b_0(t))\delta(t - \xi) \quad (14)$$

subject to the initial condition

$$\frac{\partial^\nu h(t, \xi)}{\partial t^\nu}\Big]_{t=\xi-} = 0, \qquad \nu = 0, 1, \cdots, n-1.$$

(For a more detailed discussion of this and related problems, see Borsky [6], Solodov [5], and Zadeh [4].

[1] It is generally unrecognized that a differential equation such as (12) does not characterize a system uniquely unless it is tacitly understood that the system must be nonanticipative. More specifically, (12) defines $n+1$ nonequivalent systems of which one is nonanticipative, one is purely anticipative, and the rest operate on both the past and the future of the input.

Usually, the problem is to pass from an implicit characterization employing differential operators to an explicit characterization involving such functions as the impulsive response, frequency-response function, etc. In some cases, however, one is faced with a converse problem, namely, that of passing from, say, the impulsive response to the differential equation. This may happen, for example, when it is desired to simulate a system on an analog computer which is designed to handle differential equations.

An interesting special case of the latter type of problem was considered by White [7]. Specifically, White considers the case where one is given the eigenfunction set $\{k(t, \lambda)\}$ of an undetermined differential operator $L(p, t)$ which admits a representation of the form

$$L(p, t) = \sum_{\nu}\sum_{\mu} l_{\nu\mu}t^{\nu+\mu}p^{\nu} \quad (15)$$

where the $l_{\nu\mu}$ are unknown constants, and ν and μ range over integers. Furthermore, it is assumed that $k(t, \lambda)$ can be expanded as a power series in λt,

$$k(t, \lambda) = \sum_{n=0}^{\infty} k_n(\lambda t)^n, \quad (16)$$

and that the response function $R(\lambda)$ can likewise be expanded as

$$R(\lambda) = \sum_{s} h_s\lambda^s, \quad (17)$$

where the coefficients h_s may be adjusted to yield a simple solution to the problem if one exists.

Formally, the problem is solved by substituting (12)–(14) in

$$L(p, t)k(t, \lambda) = R(\lambda)k(t, \lambda), \quad (18)$$

and equating the coefficients of the like powers of λ and t. This yields a set of equations

$$\sum_{\nu} l_{\nu,-\mu}\frac{n!}{(n-\nu)!} = h_\mu\frac{k_{n-\mu}}{k_n}, \quad (19)$$

which may be solved for the $l_{\nu\mu}$ (if a solution exists).

For example,

$$k(t, \lambda) = J_0(\lambda t) \qquad (\textit{B}\text{essel function}) \quad (20)$$

$$= \sum_{n=0}^{\infty}(-1)^{n/2}\frac{(\lambda t)^n}{2^n[(n/2)!]^2} \qquad n = 0, 2, 4, \cdots. \quad (21)$$

In this case, it is expedient to set $h_2 = -1$, $h_0 = h_1 = h_3 = \cdots = 0$. Then

$$\sum_{\nu} l_{\nu,-\mu}\frac{n!}{(n-\nu)!} = -1[-4(n/2)^2], \quad (22)$$

or

$$+l_{0,-2} + l_{1,-2}n + l_{2,-2}n(n-1) + \cdots = n^2 \quad (23)$$

from which it follows that

$$l_{1,-2} = 1, \qquad l_{2,-2} = 1, \qquad \text{all other } l_{\nu\mu} = 0.$$

Thus, the desired operator is

$$L(p, t) = p^2 + \frac{1}{t} p. \tag{24}$$

The relation given by (19) can readily be extended to the case where $k(t, \lambda)$ admits a more general representation of the form

$$k(t, \lambda) = \sum_n \sum_m k_{nm} \lambda^n t^m. \tag{25}$$

Then (19) becomes

$$\sum_m \sum_\nu k_{nm} l_{\nu, \mu - m} \frac{m!}{(m - \nu)!} = \sum_m h_{n-m} k_{m\mu}, \tag{26}$$

which, for the special case of (16), reduces to (19).

If the eigenfunction set is identified with the functions $\{t^{-\lambda}, 0 < t < \infty\}$, with λ taking values on a Bromwich-Wagner contour in the λ plane, then the differential operators relating y to u assume the form

$$(a_n t^n p^n + a_{n-1} t^{n-1} p^{n-1} + \cdots + a_0) y = (b_m t^m p^m + \cdots + b_0) u, \tag{27}$$

where the a's and b's are constants. Networks described by equations of this form have been studied by Aseltine [8], Aseltine and Trautman [9], Davis [10], Ho and Davis [11], Gerardi [12], and others, and are referred to in the literature as the Euler-Cauchy networks. In effect, such networks can be obtained from time-invariant networks by the transformation of time-scale $t \rightarrow -\log t$. Based on this observation, it can readily be shown that the impulsive response of a non-anticipative Euler-Cauchy network is of the form

$$h(t, \xi) = 1(t - \xi) \frac{1}{t} g\left(\frac{t}{\xi}\right) \tag{28}$$

where $g(t-\xi)$ is the impulsive response of the time-invariant network from which the Euler-Cauchy network is derived.

If $k(t, \lambda)$ is identified with $1/(2\pi j)(t^{-\lambda})$, then the inverse kernel is given by

$$k^{-1}(\lambda, t) = t^{\lambda - 1}, \qquad 0 \leq t < \infty. \tag{29}$$

Thus, the resolution of a signal $u(t \geq 0)$ into the $k(t, \lambda)$ assumes the form

$$u(t) = \frac{1}{2\pi j} \int_C t^{-\lambda} U(\lambda) d\lambda \tag{30}$$

and

$$U(\lambda) = \int_0^\infty u(t) t^{\lambda - 1} d\lambda \tag{31}$$

where $U(\lambda)$ is the Mellin transform of u. In some cases, it is more convenient to employ a modified Mellin transform of u which is related to (31) by the translation $t \rightarrow t + a$, where a is a constant. Such transforms have been used for illustrative purposes by Davis [10]. More recently, Peschon [13] has found a useful application for modified Mellin transforms in final value control problems.

An interesting idea which was motivated by the work of Aseltine and Trautman is contained in the paper by Davis [10] cited above. Specifically, let

$$L = a_1(t) p + a_0(t) \tag{32}$$

be a first-order differential operator and let $k(t, \lambda)$ and $k^{-1}(t, \lambda)$ be, respectively, the solutions of

$$Ly = \lambda y, \qquad y(0) = 1 \tag{33}$$

and

$$L^* y = \lambda y, \qquad y(0) = 1 \tag{34}$$

where

$$L^* = - p a_1(t) + a_0(t) \tag{35}$$

is the adjoint[2] of L. Then the kernels $k(t, \lambda)$ and $k^{-1}(\lambda, t)$ are inverse to one another in the sense of (6), and, under mildly restrictive conditions, one can write

$$U(\lambda) = \int_0^\infty k^{-1}(t, \lambda) u(t) dt \tag{36}$$

and

$$u(t) = \frac{1}{2\pi j} \int_C k(t, \lambda) U(\lambda) d\lambda. \tag{37}$$

Now, given a time-invariant network N, one can transform it into a time-varying network N^* by replacing each unit inductor (which corresponds to the operator p) with a network l corresponding to operator L, and replacing each unit capacitor with a network l' corresponding to the operator L^{-1} (inverse of L), leaving all resistances unchanged. The networks l and l' corresponding to L and L^{-1}, respectively, are described by the voltage-current relations $v = Li$ and $i = Lv$. Thus, l' can be obtained from l by dualization. Clearly, if a network function associated with time-invariant N is a function $H(s)$ of complex-frequency s, then the corresponding response function for time-varying N^* will be $H(\lambda)$, with λ being the parameter entering into $k(t, \lambda)$. In this way, one can derive from a prototype time-invariant network N a wide variety of time-varying networks N^* which are described by the same equations

[2] The adjoint of a differential operator

$$L = \sum_n a_n(t) p^n$$

is defined as

$$L^* = \sum_n (-1)^n p^n a_n(t),$$

where $p^n a_n(t)$ signifies that p^n operates on the product of $a_n(t)$ and the operand; *e.g.*, $(pa_1(t) + a_0(t))u(t)$ means $d/(dt)(a_1(t)u(t)) + a_0(t)u(t)$. For a more detailed discussion see, for instance, Ince [14], Friedman [15], and Laning and Battin [16]. It should be noted, alternatively, that the adjoint of a system B with impulsive response $h(t, \xi)$ can be defined as a system B^* whose impulsive response $h^*(t, \xi)$ is $h(\xi, t)$, *i.e.*, $h^*(t, \xi) \equiv h(\xi, t)$. With this as the starting point, it is a simple matter to show that if B is characterized by the differential equation $Ly = u$, then B^* is characterized by the differential equation $L^* y = u$, where L^* is the adjoint of L in the sense defined above. However, if B is characterized by $Ly = Mu$, then, except in special cases, the characterizing equation for B *is not* $L^* y = M^* u$.

in terms of λ as N is in terms of s. In effect, Davis' method amounts to an extension of the familiar frequency-transformation technique (see Laurent [17], Bode [17a], and Zadeh [17b]), to time-varying networks. More recently, somewhat overlapping results have been obtained independently by Wattenburg [18].

There are several other contributions, notably by Huggins [19], Lai [20], Gerlach [21], Narendra [22], and others, which are concerned with various techniques of resolution of time functions into elementary signals. These contributions will be discussed in Part II of the paper.

Transition from Impulsive Response to the Differential Equation

The problem of finding the differential equation of a system which is characterized by its impulsive response arises in situations where it is simpler to simulate or synthesize a system in terms of its differential equations rather than its impulsive response. It arises also in connection with the so-called factorization problem which is discussed in Section III of this paper.

This problem was first formulated and partially solved by Batkov [23]. Batkov's approach is based on the well-known fact (see, for instance, Coddington and Levinson [24], and Miller [3]) that the impulsive response of a system characterized by the differential equation

$$Ly = (a_n(t)p^n + \cdots + a_0(t))y(t) = u(t) \tag{38}$$

has the form

$$h(t, \xi) = 1(t - \xi) \sum_{\nu=1}^{n} \psi_i(t)\psi_i^*(\xi) \tag{39}$$

where the $\psi_i(t)$ are any n linearly independent solutions of the homogeneous equation $Ly=0$ and the $\psi_i^*(t)$ are n linearly independent solutions of the adjoint equation

$$L^*y = ((-1)p^n a_n(t) + \cdots + a_0(t))y(t) = 0. \tag{40}$$

The $\psi_i^*(t)$ can be expressed in terms of the $\psi_i(t)$ by solving the system of n linear equations in the $\psi_i^*(\xi)$

$$\left.\frac{\partial^k h(t, \xi)}{\partial t^k}\right]_{t=\xi} = 0, \qquad k = 0, 1, \cdots, n-2 \tag{41}$$

$$\left.\frac{\partial^{n-1} h(t, \xi)}{\partial t^{n-1}}\right]_{t=\xi} = \frac{1}{a_n(\xi)}. \tag{42}$$

Another well-known fact is that the operator L can readily be constructed from the knowledge of the $\psi_i(t)$. Specifically (see, for example, Ince [14]), one can write

$$Ly = \begin{vmatrix} \psi_1(t) & \cdots & \psi_n(t) & y(t) \\ \psi_1^{(1)}(t) & \cdots & \psi_n^{(1)}(t) & y^{(1)}(t) \\ \cdots & & & \\ \psi_1^{(n)}(t) & & \psi_n^{(n)}(t) & y^{(n)}(t) \end{vmatrix}. \tag{43}$$

Thus, if $h(t, \xi)$ is given in the form (39) and it is known that the system is characterizable in the form (1), then L can be obtained at once from (43). Far less trivial is the more general case where the differential equation characterizing the system is of the form $Ly = Mu$ or, more explicitly,

$$[a_n(t)p^n + \cdots + a_0(t)]y = [b_m(t)p^m + \cdots + b_0(t)]u, \tag{44}$$

where $m < n$ and the $a_i(t)$ and $b_j(t)$ are initially unknown coefficients.[3] In this case, the impulsive response reads

$$h(t, \xi) = 1(t - \xi) \sum_{i=1}^{n} \psi_i(t)\theta_i(\xi) \tag{45}$$

where the $\psi_i(t)$, $i=1, \cdots, n$ are n linearly independent solutions of $Ly=0$ and the $\theta_i(t)$, $i=1, \cdots, n$, are given by

$$\theta_i(t) = M^*[\psi_i^*(t)], \qquad i = 1, \cdots, n, \tag{46}$$

where the $\psi_i^*(t)$ are n linearly independent solutions of the adjoint equation $L^*y=0$ (satisfying [41] and [42]) and M^* is the adjoint of M. If $h(t, \xi)$ is given in this form, then from (45) one can find the $\psi_i(t)$ and $\theta_i(t)$. Once the $\psi_i(t)$ are known, it is a relatively simple matter to determine the $\psi_i^*(t)$.[4] Thus, the problem reduces to finding M^* from (46). On writing

$$M^* = b_m^*(t)p^m + \cdots + b_0^*(t) \tag{47}$$

where the $b_i^*(t)$ are undetermined coefficients, (46) yields a set of n simultaneous equations in the $b_i^*(t)$

$$\theta_i(t) = \sum_{k=0}^{n-1} b_k^*(t)\psi_i^{*(k)}(t), \qquad i = 1, \cdots, n. \tag{48}$$

Once these n equations are solved for the $m+1$ coefficients $b_k^*(t)$ in M^*, M can be found from the relation between a differential operator and its adjoint.[2] This completes the determination of L and M in (44) from the knowledge of $h(t, \xi)$.

It is simpler and frequently more effective to derive from $h(t, \xi)$, not a single differential equation of the form $Ly = Mu$, but a system of n first-order differential equations of the form

$$\dot{\boldsymbol{x}}(t) = \boldsymbol{A}(t)\boldsymbol{x}(t) + \boldsymbol{B}(t)u(t) \tag{49}$$

together with the expression for the output

$$y(t) = \alpha_1(t)x_1(t) + \cdots + \alpha_n(t)x_n(t) + \beta_0(t)u(t) + \cdots + \beta_{m-n}(t)u^{(m-n)}(t) \tag{50}$$

[3] The procedure for finding the operator M given in Batkov's paper is somewhat less straightforward than that given here.

[4] See Ince [14], p. 124.

where $\mathbf{x}(t)=(x_1(t), \cdots, x_n(t))$ is a state vector, $\boldsymbol{A}(t)$ and $\boldsymbol{B}(t)$ are matrices, and the $\alpha_i(t)$ and $\beta_k(t)$ are scalar coefficients. These equations constitute a very useful mode of characterization of a system and will be referred to as its state equations. For a homogeneous equation $Ly=0$, the state equation $\dot{\mathbf{x}}=\boldsymbol{A}\mathbf{x}$ is simply a normal form of $Ly=0$ (see, for instance, Ince [14]). Equations of the form (49) and (50) have been employed extensively in analytical dynamics and, more recently, in the theory of automatic control.

We shall discuss briefly three problems arising in connection with the state equations of a system: first, the problem of obtaining the representation (49) and (50) from $h(t, \xi)$; second, the problem of obtaining (49) and (50) from $Ly=mu$; and third, the problem of obtaining (49) and (50) directly from the structure of a time-varying network.

Batkov [23] and, somewhat earlier, Darlington [25] have indicated how the first two of these problems can be solved. As for the third problem, Kinarawala [26] has shown how one can adapt Bashkow's A-matrix technique [27] to time-varying networks and thereby obtain (49) for any $R(t)L(t)C(t)$ network without the necessity of performing differentiations or integrations on the loop or node equations of the network.

The basic idea behind the methods of Darlington and Batkov is closely related to the partial-fractions technique of Lur'e [28][5] for casting into the diagonal or, more generally, the Jordan canonical form, the dynamical equations of the linear part of a nonlinear system. Specifically, we observe that each term in the expression for $h(t, \xi)$

$$h(t, \xi) = 1(t-\xi)\sum_{i=1}^{n}\psi_i(t)\theta_i(\xi) \tag{51}$$

can be identified with the impulsive response of a first-order system. Thus, we can write

$$y(t) = x_1(t) + \cdots + x_n(t) \tag{52}$$

where $x_i(t)$ is defined by

$$x_i(t) = \int_{-\infty}^{t}\psi_i(t)\theta_i(\xi)u(\xi)d\xi. \tag{53}$$

Differentiating both sides of (53), we get

$$\dot{x}_i(t) = \psi_i(t)\theta_i(t)u(t) + \int_{-\infty}^{t}\dot{\psi}_i(t)\theta_i(\xi)u(\xi)d\xi \tag{54}$$

which can be written as

$$\dot{x}_i(t) = \frac{\dot{\psi}_i(t)}{\psi_i(t)}x_i(t) + \psi_i(t)\theta_i(t)u(t), \quad i=1, \cdots, n. \tag{55}$$

[5] See also B. V. Bulgakov, "On normal coordinates," *Prikl. Mat. i Meh.*, vol. 10, no. 2, pp. 273–290; 1946.

These equations, together with (52), form the state equations for the system, with

$$A = \begin{bmatrix} \dot{\psi}_1/\psi_1 & 0 & 0 & \cdot & 0 \\ 0 & \dot{\psi}_2/\psi_2 & 0 & \cdot & 0 \\ 0 & 0 & 0 & \cdot & \dot{\psi}_n/\psi_n \end{bmatrix},$$

$$B = \begin{bmatrix} \psi_1\theta_1 \\ \psi_2\theta_2 \\ \psi_n\theta_n \end{bmatrix} \tag{56}$$

and $\alpha_1=\cdots=\alpha_n=1$, $\beta_0=\cdots=\beta_{m-n}=0$ (assuming that $m<n$). The main advantage of this technique is that it yields a diagonal A matrix.[6]

The same idea can be used to derive an analog of the partial-fractions expansion for time-varying networks. As an illustration, consider a second-order system characterized by a single differential equation

$$(p^2 + a_1(t)p + a_0(t))y = (b_1(t)p + b_0(t))u \tag{57}$$

and let ψ_1 and ψ_2 be any two linearly independent solutions of $(p^2+a_1(t)p+a_0(t))y=0$. Then, on denoting $\dot{\psi}_1/\psi_1$ by λ_1, $\dot{\psi}_2/\psi_2$ by λ_2, and using the results just described, we can write (with the dependence of x_1, λ_1, ψ_1, etc., on t implied but not exhibited explicitly)

$$\dot{x}_1 = \lambda_1 x_1 + \psi_1\theta_1 u \tag{58}$$

$$\dot{x}_2 = \lambda_2 x_2 + \psi_2\theta_2 u \tag{59}$$

$$y = x_1 + x_2 \tag{60}$$

which is equivalent to expressing y in the form

$$y = [(p-\lambda_1)^{-1}\psi_1\theta_1]u + [(p-\lambda_2)^{-1}\psi_2\theta_2]u \tag{61}$$

where $(p-\lambda_1)^{-1}$ and $(p-\lambda_2)^{-1}$ denote the operators inverse to $p-\lambda_1$ and $p-\lambda_2$, respectively. In effect, this amounts to the partial-fractions expansion

$$(p^2 + a_1p + a_0)^{-1}(b_1p + b_0) = (p-\lambda_1)^{-1}\psi_1\theta_1 + (p-\lambda_2)^{-1}\psi_2\theta_2 \tag{62}$$

where, to recapitulate, ψ_1, ψ_2 are two linearly independent solutions of $(p^2+a_1p+a_0)y=0$

$$\lambda_1 = \dot{\psi}_1/\psi_1, \qquad \lambda_2 = \dot{\psi}_2/\psi_2 \tag{63}$$

$$\theta_1 = (-pb_1 + b_0)\psi_1^*, \qquad \theta_2 = (-pb_1 + b_0)\psi_2^* \tag{64}$$

and ψ_1^* and ψ_2^* are two linearly independent solutions of the adjoint equation $L^*y=(p^2-pa_1+a_0)y=0$ (satisfying [41] and [42]).

Partial-fractions expansions of this type and their application to the analysis and synthesis of time-varying networks have their origin in the work of Darlington [25], [29], [30].

[6] This technique results in a diagonal A even in the case of a time-invariant system whose characteristics equation $L(p)=0$ has repeated roots. However, the price for this is that one or more of the elements of A are functions of time rather than constants. If it is required that the elements of A be constants, then A can always be put into the Jordan canonical form but not, in general, the diagonal form.

The methods described above furnish a straightforward way for effecting a transition from a single differential equation

$$Ly = u, \qquad (L \text{ of order } n) \tag{65}$$

to the state equations

$$\dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}u, \qquad \mathbf{A} \text{ diagonal} \tag{66}$$

$$y = \alpha_1 x_1 + \cdots + \alpha_n x_n \tag{67}$$

provided one knows a set of n linearly independent solutions of the homogeneous equation $Ly=0$. Note that if $\mathbf{A}$ is not required to be diagonal, then the transition can be effected in conventional ways without the knowledge of solutions of $Ly=0$.

From (66) and (67), one can pass to a more explicit characterization

$$\mathbf{x}(t) = \mathbf{G}(t, t_0)\mathbf{x}(t_0) + \int_{t_0}^{t} \mathbf{G}(t, \xi)\mathbf{B}u(\xi)d\xi \tag{68}$$

$$y = \alpha_1 x_1 + \cdots + \alpha_n x_n \tag{69}$$

by using the standard form of solution of the vector equation (66). (See, for instance, Coddington and Levinson [24] and Laning and Battin [16].) The $n \times n$ matrix $\mathbf{G}(t, t_0)$ appearing in (68) has as its ith column $(i=1, \cdots, n)$ the functions $\psi_i(t, t_0)$, $\psi_i^{(1)}(t, t_0), \cdots, \psi^{(n-1)}(t, t_0)$ where $\psi_i(t, t_0)$ is the solution of $Ly=0$ which satisfies the initial conditions

$$\left.\frac{\partial \psi_i^{(k)}(t, t_0)}{\partial t^k}\right]_{t=t_0} = 0$$

for $k \neq i-1$ (with k ranging over the set $0, \cdots, n-1$) and

$$\left.\frac{\partial \psi_i^{(k)}(t, t_0)}{\partial t^k}\right]_{t=t_0} = 1$$

for $k=i-1$. $\mathbf{G}(t, t_0)$ is known by a variety of names of which the *transition matrix* is the more frequently used in engineering literature. (For a more detailed discussion of the properties of $\mathbf{G}(t, t_0)$, see, for instance, Laning and Battin [34] and Kalman and Bertram [35].)

Now suppose that one does not know a set of n linearly independent solutions of $Ly=0$ or $\dot{\mathbf{x}}=\mathbf{A}\mathbf{x}$. In these circumstances, there is no general way of passing from (65) or (66) to (68). However, there are special cases for which $\mathbf{G}(t, t_0)$ can readily be found. One such case has been discussed by Gauthier [36] and independently by Kinarawala [26]. Specifically, this is the case where, as in the constant coefficient case, $\mathbf{G}(t, t_0)$ can be expressed as

$$\mathbf{G}(t, t_0) = \exp - \int_{t_0}^{t} \mathbf{A}(\xi)d\xi. \tag{70}$$

It is easy to show that $\mathbf{G}(t, t_0)$ can be written in this form if, and only if, the matrices $\mathbf{A}(t)$ and $\int_{t_0}\mathbf{A}(\xi)d\xi$ commute, *i.e.*,

$$\mathbf{A}(t)\int_{t_0}^{t}\mathbf{A}(\xi)d\xi \equiv \left(\int_{t_0}^{t}\mathbf{A}(\xi)d\xi\right)\mathbf{A}(t). \tag{71}$$

We turn next to an important problem on which the techniques discussed in this section have a significant bearing. This is the problem of factorization.

III. Problem of Factorization

In dealing with stationary as well as nonstationary time series and, more particularly, with problems of prediction, detection and filtering, it is frequently expedient to employ so-called spectrum-shaping techniques which allow one to modify at will the spectral density and the correlation function of a random process. The spectrum-shaping technique was first used by Bode and Shannon [31] and independently by Zadeh and Ragazzini [32]. Its extension to time-varying systems was given by Laning and Battin [16] and Miller and Zadeh [33].

In the case of nonstationary processes, a typical problem in spectrum shaping is the following. Given a nonstationary process $\{u(t), -\infty < t < \infty\}$ with covariance function $R(t, \tau)$, find a linear network B such that the covariance function of the process resulting from acting on $\{u\}$ with B is $\delta(t-\tau)$ (or, more generally, a linear combination of δ functions of various orders). In other words, we wish to find a B such that $\{y(t), -\infty < t < \infty\}$, where $y=Bu$, has white spectrum. Furthermore, we wish to characterize B by a single differential equation.

This problem has an important bearing on the solution of an integral equation which is encountered very frequently in the prediction, detection and filtering of nonstationary processes. The equation in question reads

$$f(t) = \int_{a}^{b} R(t, \tau)x(\tau)d\tau, \qquad a \leq t \leq b \tag{72}$$

where $R(t, \tau)$ is a covariance function, $f(t)$ is a given function on $[a, b]$, and $x(t)$ is an unknown function. It is easy to show (see, for instance, Miller and Zadeh [33]) that the solution of this equation can be reduced to the factorization of $R(t, \tau)$. In this connection, it is of interest to note that the methods given by Laning and Battin [16], Zadeh and Miller [33], Shinbrot [37], [38], Pugachev [39], [40], and others, for the solution of this equation are either explicitly based on the factorization of $R(t, \tau)$ or make an implicit use of it.

For our purposes, it will be somewhat more convenient to discuss a converse problem, namely, that of finding a spectrum-shaping network B such that the covariance

function of the process resulting from acting with B on white noise is $R(t, \tau)$. If the impulsive response of B is denoted by $h(t, \xi)$, then it is easy to verify that

$$R(t, \tau) = \int_{-\infty}^{\infty} h(t, \xi)h(\tau, \xi)d\xi. \tag{73}$$

Thus, in this case the factorization problem becomes that of solving the integral equation (73) for $h(t, \xi)$. [Note that if $h(t, \xi)$ is assumed to be nonanticipative, then the upper limit in (73) becomes Min (t, τ).]

If B is characterizable by a differential equation of order n, then its impulsive response is of the form (45)

$$h(t, \xi) = 1(t - \xi) \sum_{i=1}^{n} \psi_i(t)\theta_i(\xi) \tag{74}$$

where the meaning of the ψ_i and θ_i was discussed previously. On substituting this expression into (73), one gets, after some straightforward manipulations

$$R(t, \tau) = \sum_{i=1}^{n} \psi_i(t)\gamma_i(\tau), \qquad t > \tau \tag{75}$$

$$= \sum_{i=1}^{n} \psi_i(\tau)\gamma_i(t), \qquad t < \tau \tag{76}$$

where

$$\gamma_i(t) = \sum_{j=1}^{n} \psi_j(t) \int_{-\infty}^{t} \theta_i(\xi)\theta_j(\xi)d\xi, i = 1, \cdots, n.$$

Thus, if $R(t, \tau)$ is given in this form, then the problem of finding the differential equation can be solved by first finding $h(t, \xi)$ [based on (74)–(76)] and then using $h(t, \xi)$ to determine the differential equation by the techniques described in the preceding section. Essentially, this approach to the determination of the spectrum-shaping network was developed by Darlington [29] and, independently, by Batkov [41]. Darlington, in particular, has considered the questions of existence and uniqueness of a nonanticipative (physically realizable) impulsive response $h(t, \xi)$ satisfying the integral euqation (73). However, neither Darlington nor Batkov have proved that any given real-valued covariance function (satisfying the conditions of symmetry and positive definiteness) can be expressed in the form (73), with $h(t, \xi)$ being a real-valued function vanishing for $\xi > t$.

As is well known for the time-invariant case, a rational spectral-density function $S(\omega)$ [$S(\omega)$ = Fourier transform of $R(t)$, with $R(t) = R(t, 0)$ = autocorrelation function] can be expressed in an infinity of ways as a product of factors $H_1(j\omega)$ and $H_2(j\omega)$

$$S(\omega) = H_1(j\omega)H_2(j\omega) \tag{76a}$$

if no restrictions are imposed on $H_1(j\omega)$ and $H_2(j\omega)$ other than that H_1 and H_2 be rational and real functions of $j\omega$. However, if 1) $H_1(j\omega)$ is required to be minimum phase (no poles or zeros in the right-half plane), which is equivalent to requiring that the impulsive responses $h_1(t)$ and $h_1^{-1}(t)$, corresponding to $H_1(j\omega)$ and $1/H_1(j\omega)$, respectively (h_1 = inverse Fourier transform of H_1, h_1^{-1} = inverse Fourier transform of H_1^{-1}), be nonanticipative, and 2) $H_2(j\omega)$ is required to be maximum phase (no poles or zeros in the left-half plane), which is equivalent to requiring that h_2 and h_2^{-1} be purely anticipative, *i.e.*, $h_2(t) = h_2^{-1}(t) = 0$ for $t > 0$, then the factorization (76a) is essentially unique.

Now in the time-varying case, the algebraic decomposition (76a) which corresponds to the convolution

$$R(t) = \int_{-\infty}^{\infty} h_1(t - \xi)h_2(\xi)d\xi \tag{76b}$$

is replaced by

$$R(t, \tau) = \int_{-\infty}^{\infty} h_1(t, \xi)h_2(\xi, \tau)d\xi \tag{76c}$$

which is the composition[7] of impulsive responses h_1 and h_2. Equivalently, (76c) may be written as

$$R(t, \tau) = \int_{-\infty}^{\infty} h_1(t, \xi)h_2^*(\tau, \xi)d\xi \tag{76d}$$

where h_2^* is the adjoint of h_2 [*i.e.*, $h_2^*(t, \xi) = h_2(\xi, t)$].[2] Thus, in the time-varying case, the algebraic problem of decomposition of $S(\omega)$ into two factors $H_1(j\omega)$ and $H_2(j\omega)$ satisfying 1) and 2) becomes that of solving the integral equation (76c) for h_1 and h_2, under the condition that $h_1(t, \xi)$ and $h_1^{-1}(t, \xi)$ vanish for $t < \xi$, and $h_2(t, \xi)$ and $h_2^{-1}(t, \xi)$ vanish for $t > \xi$.

If h_1 and h_2 satisfy these conditions and $R(t, \tau)$ is symmetric in its arguments [which in the stationary case corresponds to the condition that $S(\omega)$ is an even function of ω], then $h_1 = h_2^*$ and (76d) becomes

$$R(t, \tau) = \int_{-\infty}^{\text{Min}(t,\tau)} h(t, \xi)h^*(\xi, \tau)d\xi \tag{76e}$$

which is equivalent to (73). Now if: a) $h(t, \xi)$ is regarded as the impulsive response of a system characterized by a differential equation

$$Ly = Mu \tag{76f}$$

in which L and M are undetermined differential operators, and b) $R(t, \xi)$ is given in the form (75), (76), then by using the techniques of Darlington and Batkov it is

[7] The composition of $h_1(t, \xi)$ and $h_2(t, \xi)$ is defined as

$$h_3(t, \xi) = \int_{-\infty}^{\infty} h_1(t, \lambda)h_2(\lambda, \xi)d\lambda.$$

Essentially, if B_i is a system whose impulsive response is $h_i (i = 1, 2, 3)$, then h_3 is the impulsive response of B_1B_2, that is, the tandem combination of B_1 and B_2, with B_1 operating on the output of B_2.

a relatively simple matter to determine L. The determination of M, however, reduces to a problem for which neither Darlington nor Batkov give explicit solutions except in certain special cases. Specifically, the problem is that of finding a representation for a given self-adjoint differential operator $K(K=K^*)$ in the form

$$K = M(p, t)M^*(p, t) \tag{76g}$$

where the product should be understood in the operator sense. In the time-invariant case, this problem reduces to the straightforward algebraic problem of factoring an even polynomial in p into the product of two real polynomials which are complex conjugates of one another. In the time-varying case, M and M^* can readily be determined if one knows a set of basis functions[8] for K (which is analogous to knowing the zeros of K in the time-invariant case). Unfortunately, the assumption that K is a differential operator with a known set of basis functions is rarely satisfied in practice.

A different and more general approach to the factorization problem was developed by Levy in a series of papers [42a]–[42c] dealing with the theory of Wiener processes.[9] A fairly detailed exposition of those aspects of Levy's method which are relevant to the factorization problem are contained in the appendix of Batkov's paper [41]. Briefly, let $\eta(t, \tau)$ and $\beta(\theta, \lambda)$ denote the expressions appearing on the right side of (77) and (78):

$$\eta(t, \tau) = \frac{\dfrac{\partial^{2n-2m}R(t, \tau)}{\partial t^{n-m}\partial \tau^{n-m}}}{\left[\dfrac{\partial^{n-m}h(t, \lambda)}{\partial t^{n-m-1}}\right]_{\lambda=t}\left[\dfrac{\partial^{n-m-1}h(t, \lambda)}{\partial t^{n-m-1}}\right]_{\lambda=\tau}} \tag{77}$$

$$\beta(\theta, \lambda) = \frac{\dfrac{\partial^{n-m}h(\theta, \lambda)}{\partial \theta^{n-m}}}{\left[\dfrac{\partial^{n-m-1}h(\theta, \lambda)}{\partial \theta^{n-m-1}}\right]_{\lambda=\theta}}, \qquad \theta > \lambda \tag{78}$$

it is easy to verify that $\beta(t, \xi)$ satisfies a Volterra equation of the second kind

$$\eta(t, \tau) = \int_0^t \beta(t, \lambda)\beta(\tau, \lambda)d\lambda + \beta(t, \tau), \qquad t < \tau. \tag{79}$$

The resolvent of this equation, $\alpha(t, \xi)$, is by definition a function satisfying the integral equation

$$\beta(t, \xi) = \alpha(t, \xi) + \int_0^t \beta(\lambda, \xi)\alpha(t, \lambda)d\lambda. \tag{80}$$

From this and the preceding equations, it can be deduced that $\alpha(t, \xi)$ satisfies a Fredholm integral equation of the second kind

$$\alpha(t, \xi) = \eta(t, \xi) - \int_0^t \eta(\lambda, \xi)\alpha(t, \lambda)d\lambda, \qquad t > \xi. \tag{81}$$

Now the function $\eta(t, \xi)$ in (81) can be expressed explicitly in terms of $R(t, \tau)$ and its derivatives by expressing the two factors in the denominator of $\eta(t, \xi)$ [see (77)] in terms of $R(t, \tau)$. For example,

$$\left[\frac{\partial^{n-m-1}h(\theta, \lambda)}{\partial \theta^{n-m-1}}\right]_{\lambda=\theta} = \sqrt{(-1)^{n-m-1}R_{2n-2m-1}(\theta, \theta)}. \tag{82}$$

Thus, (81) can be solved, in principle, for $\alpha(t, \xi)$ in terms of $\eta(t, \xi)$. Then, knowing $\alpha(t, \xi)$, one can determine $\beta(t, \xi)$ from (80). Finally, $h(t, \tau)$ is found from the relation (note the interchange in arguments in h)

$$h(\tau, t) = \gamma(t, \tau) - \int_0^t \alpha(t, \lambda)\gamma(\lambda, \tau)d\lambda, \qquad t < \tau \tag{83}$$

where

$$\gamma(t, \tau) = \frac{\dfrac{\partial^{n-m}R(t, \tau)}{\partial t^{n-m}}}{\left[\dfrac{\partial^{n-m-1}h(t, \lambda)}{\partial t^{n-m-1}}\right]_{\lambda=t}}, \qquad t < \tau \tag{84}$$

and the denominator of $\gamma(t, \tau)$ is expressible in terms of $R(t, \tau)$ in the manner cited previously. In summary, by the use of this technique, the solution of the integral equation (73) for $h(t, \xi)$ is reduced to the considerably simpler problem of solving the Fredholm equation (81) for $\alpha(t, \lambda)$.

The problem of factorization is encountered also in the analysis of multipath communication systems [43], [44], where it leads to integral equations of the form (73) defined on a finite interval. Such equations have been discussed by Kailath [43] and Middleton [44].

IV. Randomly-Varying Systems

Linear randomly-varying systems play an important role in problems encountered in the study of such varied phenomena as propagation through time-varying media, reflection from fluctuating targets, turbulence, scattering, amplitude and phase modulation, magnetohydrodynamics and plasma.

Despite their importance, few if any attempts at studying the behavior of randomly-varying systems on a theoretical level were made prior to 1950. The papers published since then deal almost entirely with systems in which the random variations in parameters are stationary. This restriction is an essential one, since from a statistical viewpoint a stationary randomly-varying system behaves in some respects like a time-

[8] The basis functions of a differential operator L of order n are any n linearly independent solutions of the differential equation $Lu=0$.

[9] A Wiener process is the integral of white noise, with the limits of integration being 0 and t.

invariant system. For example, if one is observing the input and output of a time-varying black box B and it is not known whether B is linear or nonlinear, then there is no way of deciding between the two alternatives if B is nonstationary. On the other hand, if B is stationary, then it can be shown that B is linear if it has the following extended superposition property. Let $\{u(t), -\infty < t < \infty\}$ and $\{v(t), -\infty < t < \infty\}$ be two independent stationary processes which are independent also of the random processes governing the behavior of B. Let $R_u(\tau)$ and $R_v(\tau)$ be the correlation functions of the processes resulting from operating with B on the $\{u\}$ and $\{v\}$ processes, respectively. Let $R_{\alpha u+\beta v}(\tau)$ be the correlation function of the process resulting from operating with B on the process $\{\alpha u+\beta v\}$, where α and β are arbitrary real constants. If

$$R_{\alpha u+\beta v}(\tau) = \alpha^2 R_u(\tau) + \beta^2 R_v(\tau) \tag{85}$$

for all α, β, $\{u\}$ and $\{v\}$, then we shall say that B has the superposition property for correlation functions. Clearly, any stationary linear system will have this property, and the converse can also be demonstrated to be true. Consequently, (85) can be used as a basis for determining whether B is linear or nonlinear by observing the input and output of B over periods of time sufficiently long to enable the observer to obtain accurate estimates of the correlation functions involved in (85). It is tacitly understood, of course, that B is such that in the absence of input its state reverts to the ground (unexcited) state as $t \to \infty$.

The correlation function of the output y of a stationary system B which is subjected to a stationary input process $\{u\}$ can be conveniently expressed in terms of the correlation function of B. Specifically, the correlation function of a stationary randomly-varying system B is defined as follows (see Zadeh [45], [46]):

$$R(j\omega, \tau) = E\{H(j\omega, t)H(-j\omega, t+\tau)\} \tag{86}$$

where $H(j\omega, t)$ is the frequency-response function of B, and E denotes the expectation operator. (It is understood that, for each real ω, $\{H(j\omega, t), -\infty < t < \infty\}$ is a stationary random process.)

If $\{u(t)\}$ and $\{H(j\omega, t)\}$ are independent, then it can readily be shown that the correlation function of the output process, $R_y(\tau)$, is related to $R(j\omega, \tau)$ and the correlation function of the input process $R_u(\tau)$ through

$$R_y(\tau) = \frac{1}{2\pi}\int_{-\infty}^{\infty} R(j\omega, \tau)F\{R_u(\tau)\}e^{j\omega\tau}d\omega \tag{87}$$

where $F\{R_u(\tau)\}$ is the Fourier transform of the input correlation function. This relation is of the same form as the equation expressing the output of a time-varying network with frequency-response function $R(j\omega, \tau)$, with the input being $R_u(\tau)$. [Compare (3).] Relations of the form (87) have been applied by Bugnolo [47], [48] to the analysis of scattering by randomly-varying media.

In the special but significant case of a system characterized by an input-output relationship of the form

$$y(t) = u(t - \alpha(t)) \tag{88}$$

the frequency-response function reads

$$H(j\omega, t) = e^{-j\omega\alpha(t)} \tag{89}$$

where $\alpha(t)$ plays the role of a variable time delay. If $\{\alpha(t)\}$ is assumed to be a stationary Gaussian process, then the expression for $R(j\omega, \tau)$ is

$$R(j\omega, \tau) = \exp\{\omega^2[R(\tau) - R(0)]\} \tag{90}$$

where $R(\tau)$ is the correlation function of $\{\alpha(t)\}$. This result was extended by Price [50] to systems for which $H(j\omega, t)$ is of the form

$$H(j\omega, t) = \sum_i a_i e^{j\omega\alpha_i(t)} \tag{91}$$

where the $\{\alpha_i(t)\}$ are independent stationary Gaussian processes. Such systems are encountered in connection with the detection of fluctuating targets—a subject that is discussed at length by Price [51] and Price and Green [52]. A more detailed account of the results obtained by Price, Price and Green, Kailath [52], [53], and others, in connection with the identification of randomly-varying systems will be presented in Part II of this paper.

Although the correlation function $R(j\omega, \tau)$ of a stationary randomly-varying system B, or more generally, the function $R(j\omega, j\omega', \tau)$

$$R(j\omega, j\omega', \tau) = E\{H(j\omega, t)H(j\omega', t+\tau)\} \tag{92}$$

which is the expectation of the product of $H(j\omega, t)$ and $H(j\omega', t+\tau)$, conveys considerable information concerning the statistical characteristics of B, its application is limited to those systems which are characterized by their frequency-response functions. If—as is frequently the case—B is characterized by a differential equation with randomly-varying coefficients, then it is generally simpler to employ more direct techniques for studying the behavior of B in statistical terms. For example, if in

$$a_n(t)\frac{d^n y}{dt^n} + \cdots + a_0(t)y = u(t) \tag{93}$$

each $a_i(t)$ can be expressed as

$$a_i(t) = \bar{a}_i(t) + \epsilon_i(t) \tag{94}$$

where $\bar{a}_i(t)$ is the expected value of $a_i(t)$, and $\epsilon_i(t)$ is in some suitable sense small compared with $\bar{a}_i(t)$, then (93) can be solved by the usual perturbation techniques, with y expressed as the sum of a nonrandom term and

a random term due to the $\epsilon_i(t)$. [$u(t)$ is assumed to be nonrandom.] This approach and elaborations on it have been employed by, among others, Sverdrup [54], Bugnolo [47], [48], and Janos [55]. The key assumption in Janos' work, which was motivated by applications to the multipath problem, is that the coefficients $a_i(t)$ are stationary and stationarily correlated Gaussian processes. In Bugnolo's work, the perturbation technique is used mainly to yield an approximation to the correlation function of a system comprising a randomly-varying propagation medium.

An important problem in the case of a stationary randomly-varying system whose output is a nonstationary process (*e.g.*, due to the effect of initial conditions and/or the application of a nonstationary input) is that of determining the asymptotic behavior of various moments of the output as $t \to \infty$. The analysis of a first-order system

$$\dot{y} + a(t)y = u(t), \tag{95}$$

in which $a(t)$ is a randomly-varying coefficient and u is a nonrandom input, was carried out by Rosenbloom [56] for the case where $\{a(t)\}$ is a Gaussian process, not necessarily stationary. Rosenbloom considered the question of stability of such systems and noted, in particular, that under certain conditions the unit step response of a first-order randomly-varying system can tend to 1 in probability, and yet the mean value of the response may approach $-\infty$ as $t \to \infty$.

A more general analysis of first-order systems in which both $\{a\}$ and $\{u\}$ are stationary and stationarily correlated Gaussian processes was given by Tikhonov [57]. An interesting feature of Tikhonov's method is a simple and yet highly effective artifice for avoiding the difficulties arising out of the dependence of $\{a(t)\}$ and $\{u(t)\}$ processes. Specifically, the general solution of (95) reads

$$y(t) = y(t_0) \exp\left\{-\int_{t_0}^{t} a(\lambda)d\lambda\right\} + \int_{t_0}^{t} d\xi \exp\left\{-\int_{\xi}^{t} a(\lambda)d\lambda\right\} u(\xi). \tag{96}$$

In this form, the calculation of various moments of $y(t)$ is complicated by the fact that the second term on the right involves the product of

$$\exp\left\{-\int_{\xi}^{t} a(\lambda)d\lambda\right\}$$

and $u(\xi)$. These two processes are not independent and, furthermore,

$$\left\{\exp\left\{-\int_{\xi}^{t} a(\lambda)\right\} d\lambda\right\}$$

is not a Gaussian process.

As noted by Tikhonov, this apparent difficulty can be resolved by writing (96) as

$$y(t) = y(t_0) \exp\left\{-\int_{t_0}^{t} a(\lambda)d\lambda\right\} - \frac{\partial}{\partial\gamma}\left[\int_{t_0}^{t} \exp\left\{-\int_{\xi}^{t} ((\lambda)d\lambda\right\} e^{-\gamma u(\xi)}d\xi\right]_{\gamma=0} \tag{97}$$

where γ is a dummy parameter. In this expression, the product of

$$\exp\left\{-\int_{t_0}^{t} a(\lambda)d\lambda\right\}$$

and $u(\xi)$ is replaced by

$$\exp\left\{-\int_{\xi}^{t} a(\lambda)d\lambda - \gamma u(\xi)\right\},$$

which is much easier to deal with. By using this artifice, Tikhonov succeeded in obtaining an explicit expression for the correlation function of $\{y(t)\}$, which enabled him to study the asymptotic behavior of the expectation of $y(t)$ and the variance of $y(t)$ for large t.

For higher-order systems, concrete results concerning the mean-square stability were obtained by Samuels and Eringen [58], [59]. In particular, for a system of the form

$$Ly + \alpha_k(t)\frac{d^k y}{dt^k} = u(t) \tag{98}$$

in which L is a differential operator with constant coefficients, α_k is a randomly-varying parameter with correlation function $\delta(\tau)$, and u is a randomly-varying input, Samuels and Eringen obtained an explicit expression for the expectation $E\{y^2(t)\}$ as a function of t. This permitted them to determine if a system of the type under consideration is stable in mean square, that is, if

$$\lim_{t\to\infty} E\{y^2(t)\} < M,$$

where M is a finite constant. Similar results have been derived independently by Bergen [60].

On a more general level, an extension of Lyapounov's second method to randomly-varying, not necessarily linear, systems was made by Bertram and Sarachik [61]. Essentially, Bertram and Sarachik extended to systems characterized by vector differential equations of the form

$$\dot{\mathbf{x}} = f(\mathbf{x}, \omega, t) \tag{99}$$

where ω denotes an element of a probability space Ω, the basic notions of stability, asymptotic stability, asymptotic stability in the large, etc., and derived sufficient conditions *a la* Lyapounov for various types of stability. For example, the equilibrium solution

$x \equiv 0$ of (99) is said to be stable in the mean if for any $\epsilon > 0$, t_0 exists a $\delta(\epsilon, t_0) > 0$ such that for any initial x (at $t = t_0$) satisfying $\|x(t_0)\| < \delta(\epsilon, t_0)$ ($\|x\|$ denotes the norm of x), the expectation $E\{\|x(t)\|\}$ is smaller than ϵ for all $t \geq t_0$. Then—in a way paralleling that used in the second method of Lyapounov—it can be shown that if there exists a Lyapounov function $V(x, t)$ defined on the state space such that

1) $V(x, t) = 0$ for all t.
2) $V(x, t)$ is continuous in both x and t and the first partial derivatives in these variables exist.
3) For all x in the state space, $V(x, t) \geq \alpha \|x\|$ for some fixed $\alpha > 0$.
4) $E\{dV(x, t)/dt\} \leq 0$ along a solution.

Then the solution $x(t) \equiv 0$ is stable in the mean.

Unfortunately, this and related results do not provide an effective means of determining whether a given randomly-varying system is stable or unstable, since there are no general techniques for finding a Lyapounov function for a given time-varying, much less randomly-varying, system.

Indeed, the only fairly general type of randomly-varying system for which effective stability criteria have been developed is the *piecewise constant* system characterized by differential equations of the form

$$\dot{x} = A_k x, \qquad t_{k-1} \leq t < t_k, \qquad k = 1, 2, \cdots, \tag{100}$$

where x denotes the state vector, A_k is a constant matrix, and either the A_k or the t_k or both are random variables, with the A_k ranging over a finite set of constant matrices. Various special cases of such systems were studied by Bellman [62], Bellman, Harris, and Shapiro [63], Karlin [64], Kalman [65], Bergen [66], and Bharucha [67].

The principal contribution to the stability theory of such systems was made by Bellman, who was the first to point out that the Kronecker product of matrices furnishes a natural way for attacking the problem of stability of systems of this type. Specifically, Bellman considered the asymptotic behavior of products of matrices of the form

$$X_N = Z_N Z_{N-1} \cdots Z_1 \tag{101}$$

where Z_i, $i = 1, 2, \cdots, N$, are the elements of a sequence of independent, identically-distributed random matrices. For simplicity, the Z_i are assumed to range over two fixed 2×2 matrices A and B, with $\Pr\{Z_i = A\} = \Pr\{Z_i = B\} = 1/2$.

In effect, the case considered by Bellman is the discrete-time version of (100), with $t_k = k$. Thus,

$$x_k = Z_k x_{k-1}, \qquad k = 1, 2, \cdots, \tag{102}$$

and

$$x_N = (Z_N \cdots Z_1) x_0. \tag{103}$$

If the two components of x_N are denoted by x_{1N} and x_{2N}, and the norm of x is defined by

$$\|x\| = x_1^2 + x_2^2,$$

then

$$\|x_N\|^2 = x_0' Z_1' \cdots Z_N' Z_N \cdots Z_1 x_0 \tag{104}$$

where the prime signifies transposition. We are interested in the behavior of the expected value of $\|x_N\|^2$ as $N \to \infty$.

The expected value of $\|x_N\|^2$ could readily be found from (104) if the Z_i commuted with one another and their transposes, for (104) could then be written as

$$\|x_N\|^2 = x_0' Z_1' Z_1 Z_2' Z_2 \cdots Z_N' Z_N x_0, \tag{105}$$

and, on taking the expectation of both sides of (105), we would have

$$E\|x_N\|^2 = x_0' [E(Z_1' Z_1)]^N x_0 \tag{106}$$

since $E(Z_1' Z_1) = E(Z_2' Z_2) = \cdots = E(Z_N' Z_N)$, and the Z_i are mutually independent. Unfortunately, the Z_i, in general, do not commute and as a consequence the expectation of $\|x\|^2$ cannot be obtained in the simple fashion of (105) and (106).

To circumvent this difficulty, Bellman employed the notion of the Kronecker product (see, for instance, MacDuffie [68]), which is defined by the relation

$$A \otimes B = \begin{bmatrix} a_{11} b_{11} & a_{11} b_{12} & a_{12} b_{11} & a_{12} b_{12} \\ a_{11} b_{21} & a_{11} b_{22} & a_{12} b_{21} & a_{12} b_{22} \\ a_{21} b_{11} & a_{21} b_{12} & a_{22} b_{11} & a_{22} b_{12} \\ a_{21} b_{21} & a_{21} b_{22} & a_{22} b_{21} & a_{22} b_{22} \end{bmatrix} \tag{107}$$

where A and B are, for simplicity, taken to be 2×2 matrices with elements a_{ij} and b_{ij}, respectively. Similarly, the nth Kronecker power of A is defined as

$$A_{[n]} = A \otimes A \otimes \cdots \otimes A. \tag{108}$$

Thus, the second power of a column 2-vector x is given by

$$\begin{bmatrix} x_1 \\ x_2 \end{bmatrix}_{[2]} = \begin{bmatrix} x_1^2 \\ x_1 x_2 \\ x_2 x_1 \\ x_2^2 \end{bmatrix}. \tag{109}$$

A key property of the Kronecker powers of a matrix is expressed by the following identity.

$$AB_{[m]} = A_{[m]} B_{[m]} \tag{110}$$

which means that the mth Kronecker power of the product of A and B is the product of mth Kronecker powers of A and B. Based on this identity, we can derive from

$$x_k = Zk x_{k-1}, \qquad k = 1, 2, \cdots, N \tag{111}$$

the relation

$$x_{k[2]} = Z_{k[2]}x_{k-1[2]}, \qquad k = 1, \cdots, N \tag{112}$$

and hence,

$$x_{N[2]} = Z_{N[2]}Z_{N-1[2]} \cdots Z_{1[2]}x_{0[2]}. \tag{113}$$

Now we can take the expectation of both sides of (113) without the problems which were encountered in (104) as a result of the noncommutativity of the Z_i. Thus, we obtain

$$E\{x_{N[2]}\} = [E\{Z_{1[2]}\}]^N E\{x_{0[2]}\}. \tag{114}$$

Comparing $x_{N[2]}$ with $\|x\|^2$, it is clear that if $x_{N[2]} \to 0$ as $N \to \infty$, so does $\|x\|^2$, and vice versa. Consequently, the stability in the mean of the system described by (102) is governed by the behavior of $[E\{Z_{1[2]}\}]^N$ as $N \to \infty$. Specifically, if the eigenvalues of the matrix $E\{Z_{1[2]}\}$ lie inside the unit circle, then $E\{x_{N[2]}\} \to 0$ as $N \to \infty$ (assuming that $Ex_{0[2]} < \infty$), and hence, the system is asymptotically stable in the large (in the mean norm). If one or more eigenvalues lie outside the unit circle, the system is unstable in the mean norm. These and other cases are discussed in greater detail by Bharucha [67].

The case where the t_k are random variables with the $t_k - t_{k-1}$, $k = 1, 2, \cdots$, being mutually independent random variables, was first studied by Kalman [69]. Kalman, too, used Kronecker powers of matrices, but his formulas are somewhat complicated by the fact that they relate the expected value of $\|x_N\|^2$ to x_0, rather than the expected value of $x_{N[2]}$ to that of $x_{0[2]}$, as in (114).

The foregoing discussion dealt with the case of free (unforced) systems, that is, systems subjected to zero input. For forced systems, Kalman showed that if the input is bounded, then the response is bounded (in mean square) if and only if the system is stable in the mean square.

Another theorem relating the stability of the free system to the boundedness of the response to bounded inputs was given by Bharucha [69]. Bharucha's theorem is essentially an extension to randomly-varying systems of a theorem due to Perron [70] and Malkin [71]. The statement of Malkin's theorem is as follows: consider a system B characterized by the vector equation

$$\dot{x} = A(t)x + u \tag{115}$$

where x is the state vector, u is the input vector, and $A(t)$ is a bounded ($\|A(t)\| \leq M < \infty$), $0 \leq t < \infty$) and continuous matrix. Then the response of B to any bounded input is bounded if, and only if, there exist positive numbers α and β such that the norm of every solution $x(t)$ of the free system

$$\dot{x} = A(t)x, \tag{116}$$

satisfies the inequality

$$\|x(t)\| \leq \alpha e^{-\beta(t-t_0)}\|x(t_0)\|, \qquad 0 \leq t_0 < t < \infty. \tag{117}$$

In the case of a randomly-varying system, $\{A(t)\}$ is a random process. For such systems, Bharucha shows that if:

1) $\sup_t \|A(t)\|$ is almost surely bounded, and $A(t)$ is almost surely continuous.
2) $E\|G(t, \xi)\| \leq \alpha e^{-\beta(t-\xi)}$, $\alpha \geq 0$, $\beta > 0$ where $G(t, \xi)$ is the transition matrix of (115).
3) The input process is bounded and independent of $\{A(t)\}$, then $\sup_t E\|x(t))\|$ is bounded for every bounded $\|x(t_0)\|$.

In the analyses of the stability in the mean of systems defined by (102), it is usually assumed that the Z_i form a sequence of independent random matrices. This assumption is much too restrictive for many practical purposes. For example, in the case of discrete channels with fading, the successive Z_i have a high degree of dependence. As an approximation to systems of this type, it is natural to assume that the Z_i form a Markov process. The stability in the mean square of such systems was studied by Bharucha, but little if anything is known concerning their behavior when subjected to random inputs.

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Compatible Single Sideband*

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Summary—**A compatible single-sideband (CSSB) wave is a new type of modulated wave which is compatible with existing AM receivers. Spectrum analysis and measurement indicate that if the CSSB system is applied to a conventional AM broadcast transmitter, a desired-to-undesired sideband ratio of slightly better than 30 db will be achieved under normal modulation conditions. Also described is a beat frequency problem which introduces a special type of undesired sideband component. This component falls extremely close to the carrier and should not be present during the vast majority of program conditions. Thus, the technique meets the requirement of theoretically distortion-free envelope characteristics with a good desired-to-undesired sideband ratio.**

Measurements are described which show the advantages of the system. It appears that the main advantage of the technique is to reduce co- and adjacent-channel interference effects. CSSB also provides a higher fidelity signal when received by conventional inexpensive broadcast receivers. The on-the-air tests indicated good listener acceptance of the new system.

The system also appears to have applications in communications service where cost and size bar the use of conventional sideband techniques. The technique does not suffer from Doppler shift difficulties.

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Introduction

THE FACT that there is a severe shortage of spectrum space available to the broadcaster and to the high-frequency communicator has been stressed in a number of persuasive papers.[1,2]

One method of greatly easing this critical shortage is the use of the SSB. The number of SSB transmitters in operation has greatly increased during the past decade because of this serious spectrum shortage and because of a number of other operational advantages of SSB. However, there are many services that cannot justify the expense and complexity of conventional SSB equipment. These services include broadcasting and many mobile communication systems. The purpose of this paper is to describe a new type of SSB wave called compatible single sideband (CSSB), which appears to be

[1] D. G. Fink, "Danger! The radio spectrum is bursting at seams," *J. Franklin Inst.*, vol. 261, pp. 477–493; May, 1956.
[2] G. Jacobs and E. T. Martin, "The dwindling high-frequency spectrum," 1961 IRE International Convention Record, pt. 8, pp. 179–195. (To be published in IRE Trans. on Communications Systems.)

suitable for broadcasting and most mobile communications systems.

The CSSB technique discussed in this paper has recently been defined by the CCIR in the following manner: "A single-sideband transmission is considered to be compatible if it can be received on the existing conventional double-sideband receivers without any modifications whatsoever and with satisfactory quality of reception."[3]

The main advantage of CSSB over conventional SSB operation is that it may be received on the millions of conventional receivers presently in service in homes and vehicles. Equipment has been manufactured for adapting existing transmitters, and extensive tests have been run to prove that this system is completely compatible with AM transmitters as well as with AM receivers.

It might be best, right at the outset, to stress the point that the CSSB system is an altogether different system from the one proposed by the author for conventional SSB communications, that is, the envelope elimination and restoration (EER) SSB system.[4,5] The EER system, though compatible with existing class-C AM transmitters, requires the use of conventional SSB receivers and may not be received without distortion on AM receivers. The EER system was introduced in 1952 and now enjoys considerable commercial application in standard point-to-point long distance circuits.

Over the years there have been constant efforts to reduce co- and adjacent-channel interference by the use of SSB in the broadcast field. The methods proposed, while ingenious, have not received commercial acceptance. Early attempts in Great Britain and the Netherlands[6,7] suggested that double-sideband transmission be utilized for low audio-frequency components and SSB transmission be used for high audio-frequency components. Those systems depended upon the fact that, at high frequencies, only low percentages of modulation are required. Therefore, conventional SSB waves would be acceptable because, at low percentages of modulation, a conventional full carrier sideband has relatively low envelope distortion. These systems, however, do not reduce greatly co-channel interference and require replacement of many existing transmitters. Because of these limitations, these systems have not gained favor.

Recently, there have been Russian publications[8,9] indicating that intensive research has been conducted in the USSR on this same problem. They call their system "optimum modulation" but it appears that the sideband rejection of this technique is small and there is a relatively high degree of envelope distortion. It is interesting to note that Kliagin[9] provides a detailed analysis of what he calls a "Kahn transmitter." The curve of the PM wave of such a transmitter is shown in his Fig. 3. On this same figure a curve for a perfect system (but one undiscovered by Kliagin) is provided and this curve is very similar to the one described in the following discussion of the actual CSSB system.

A recent article[10] in these PROCEEDINGS described a CSSB technique developed at the University of Calcutta. The technique developed by Chakrabarti was used several years ago by this author but, because of the complexity of the equipment required for synthesizing the desired type of wave, Chakrabarti's procedure is in some respects less satisfactory than the method disclosed in this paper. It is, however, quite interesting to note that appreciable advances are being made independently at a number of research organizations throughout the world.

The first installation of equipment utilizing the CSSB technique described here was made in August, 1956, at the U. S. Voice of America megawatt station in Munich, Germany. This equipment utilized a simplified CSSB technique.

I. Problems of Utilizing the Conventional SSB Full Carrier System in AM Broadcasting Service

The reader may wonder why a SSB plus full carrier wave is not acceptable for transmission to conventional AM receivers. This section describes in some detail the reason why such two-spectral component waves are not acceptable for broadcast and high-quality AM reception.

The envelope of a carrier plus only one sideband component wave is highly distorted at high percentages of modulation. Actually, when the sideband approaches the level of the carrier, the envelope distortion is approximately 24 per cent.

Figs. 1 and 2 are curves indicating distortion for conventional SSB waves. They are taken from the section on CSSB of the *NAB Handbook* written by Harmon.[11] Fig. 1 shows the envelope function for a carrier plus SSB wave. This would be the waveshape produced at the output of a diode demodulator or any other "linear" detector in the conventional AM receiver. It should be noted that the distortion is very noticeable at high percentages of modulation.

[3] "Resolutions, Questions, and Study Programmes," Vol. 2, CCIR, 9th Plenary Assembly, Los Angeles, Calif., pp. 165–166; 1959.
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[6] P. P. Eckersley, "Asymmetric-sideband broadcasting," PROC. IRE, vol. 16, pp. 1041–1092; September, 1938.
[7] N. Koomans, "Asymmetric-sideband broadcasting," PROC. IRE, vol. 27, pp. 687–690; November, 1939.
[8] S. I. Tetal'baum and Iu. G. Grinevich, "Description of the results of experimental examination of the basic principle of the theory of optimal amplitude-phase modulation," *Radiotekhnika (Moscow)*, vol. 12; December, 1957.
[9] L. E. Kliagin, "Spectrum analysis in amplitude-phase modulation," *Radiotekhnika (Moscow)*, vol. 15, pp. 67–73; August, 1960.
[10] R. M. Golden, M. R. Schroeder and N. B. Chakrabarti, "Discussion of 'combined AM and PM for a One-Sided Spectrum'" PROC. IRE, vol. 49, pp. 1094–1095; June, 1961.
[11] R. N. Harmon, "Experience with CSSB at KDKA," in "National Association of Broadcasters Engineering Handbook," A. P. Walker, Ed., McGraw-Hill Book Co., Inc., New York, N. Y., pp. 8-41 to 8-52; 1960.

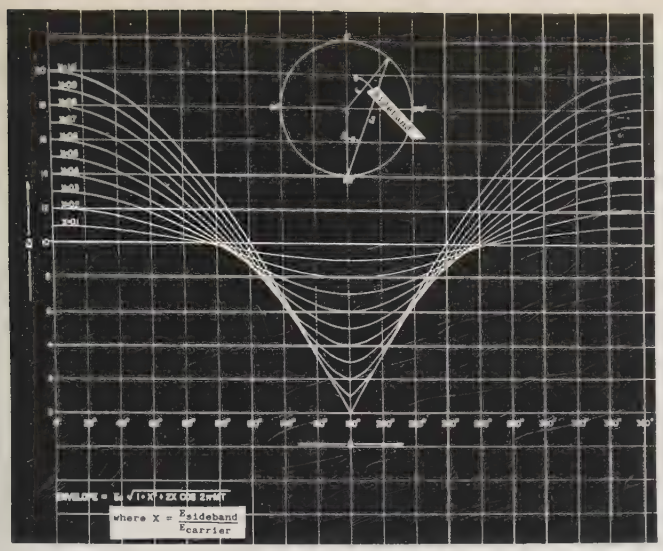


Fig. 1—Envelope wave shape of two-element carrier plus SSB wave for various ratios of sideband-to-carrier.

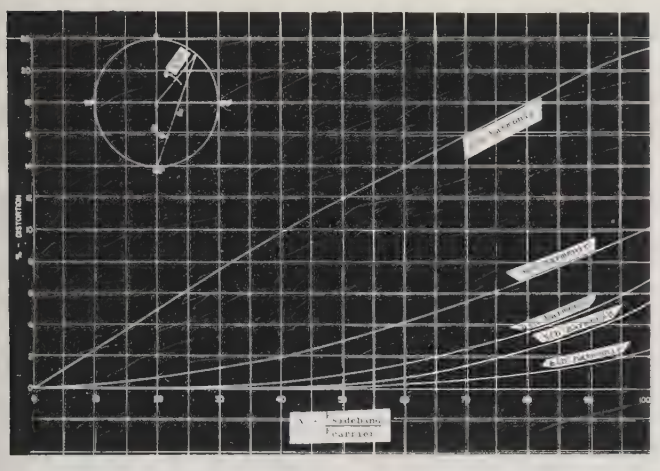


Fig. 2—Fourier series components of two-element envelope wave-shape for various ratios of sideband-to-carrier level.

Fig. 2 shows the results of a Fourier series analysis of the envelope function which was shown in Fig. 1. As mentioned before, the total harmonic distortion is approximately 24 per cent when the sideband level equals the carrier level.

Such distortion is obviously unacceptable for broadcast service and would seriously degrade even conventional communications systems. Also, it may be seen that, since the distortion accounts for much of the negative modulation swing, complete 100 per cent modulation cannot be achieved. Because of this, the maximum modulation value is equivalent to only 67 per cent of the average value of the wave. In addition, there is a severe "carrier shift;" that is, the average value of the modulated wave varies with modulation and this would slightly disturb the AVC operation of the receiver. Thus, we see that a conventional SSB wave is not suitable for high-quality broadcast use.

II. Practical Background Considerations

In analyzing any new system that claims to be compatible with existing equipment, it is necessary to understand fully the characteristics of the "existing" equipment. Careful study and experimentation with existing AM transmitters reveals that almost all AM broadcast transmitters and the vast majority of communciation transmitters are perfectly suitable for CSSB service. Generally, the transmitters are considerably better in terms of frequency response and distortion than the associated receivers. The reason for this is that the cost of the transmitter is less severely limited because generally the number of receivers far exceeds the number of transmitters. Thus, the limitation as to quality of the CSSB system, as well as the conventional AM system, is due to receiver limitations.

A number of surveys have shown that the conventional AM receiver designed for broadcast operation is severely restricted in bandwidth. A recent survey reported by a noted pioneer in radio and television research, R. B. Dome of the General Electric Company, indicated that the average broadcast receiver's bandwidth is exceedingly narrow. Dome's survey is concluded as follows:

> The 6 db bandwidths of 135 broadcast receivers indicate that some 50% of the receivers had a total IF bandwidth of less than 7.5 kc which is equivalent to less than 3.75 kc audio response at the −6 db point when conventional AM signals are received. He also reported that 79% of the receivers tested ran less than 9.5 kc total bandwidth, thus producing a high frequency −6 db cutoff point of 4.75 kc for conventional AM reception. Furthermore, his studies showed that only 9% of the receivers had a −6 db audio response to a double-sideband AM wave of 5.75 kc or higher.

It is interesting to note that these figures appear to correlate fairly well with figures provided by the British.[12]

The reason for this poor receiver frequency response is the necessity to compromise between adjacent channel interference rejection and good fidelity. Also, in inexpensive receiver design, the number of tuned circuits and IF transformers must be minimized; therefore, the shape factor of the selectivity curve generally leaves much to be desired. Fig. 3 shows an over-all fidelity characteristic of a typical home receiver of recent design.

The second basic practical fact that must be considered in an analysis of the CSSB system is the type of message to be transmitted. The CSSB system has been proposed for broadcast service and voice communications. Therefore, it is important to consider the characteristics of voice and musical waves.

It is well known that voice and musical waves are wideband signals in the sense that they cover many octaves. *However, the energy density is much higher at the*

[12] E. K. Sanderman, "Radio Engineering," Chapman and Hall Ltd., London, England, vol. 2, p. 101; 1953. See the list of characteristics of commercial receivers. Sanderman reports that the −3 db high-frequency audio cutoff point of typical sets were 1.6 kc, 3.7 kc, 3.4 kc, 5.2 kc, and 2 kc.

moderately low frequencies than at the high audio frequencies. But, even though there is a relatively small amount of energy in, say, the 3- to 5-kc range, such frequencies cannot be eliminated without loss of intelligibility. This is especially true under poor signal-to-noise ratio reception conditions. A curve has been published[13] showing the frequency characteristics of a typical musical wave; a curve showing the cumulative characteristic derived from this work is shown in Fig. 4. (A further discussion concerning typical spectra will be found in Section VII.)

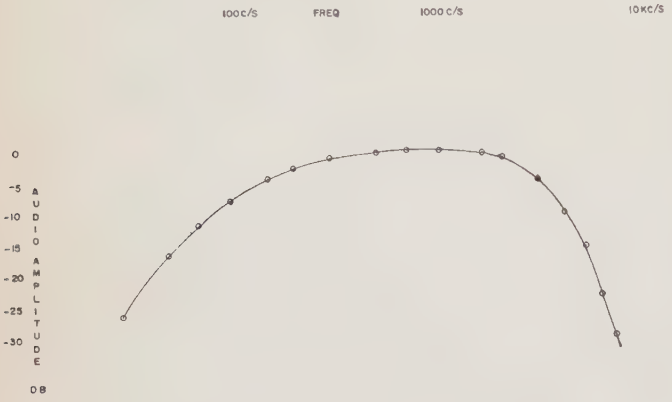


Fig. 3—Frequency response of typical broadcast receiver, RCA Model 9-C-7-EE.

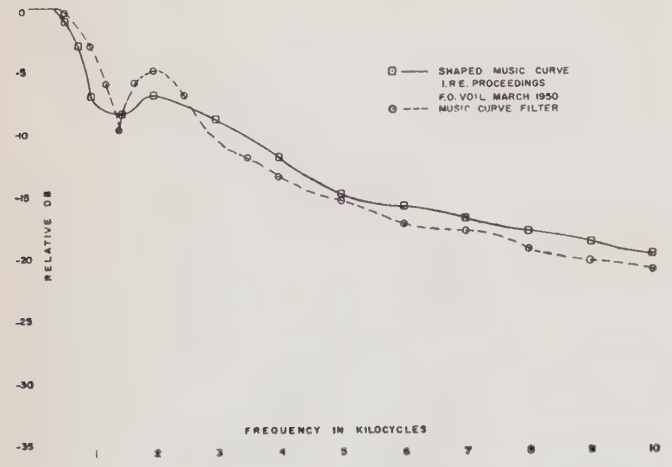


Fig. 4—Voice and music spectrum curve.

III. Basic Structure of the CSSB Wave

A perfect CSSB wave would have the following characteristics:

1) an envelope which reproduces the input modulating wave without nonlinear distortion.
2) a spectrum characteristic limited to a band extending from one side of the carrier and having a bandwidth equal to the highest audio frequency of the modulating signal.

[13] V. O. Voil, "Some problems of disk recording for broadcasting purposes," Proc. IRE, vol. 38, pp. 233–238; March, 1950.

It will be seen in the following discussion that the simultaneous satisfaction of both specifications is contradictive. But, from a practical standpoint, a very close approximation to both specifications can be and has been accomplished.

For broadcast applications, the first specification, the theoretically distortion-free envelope, is paramount. Therefore, some small liberties must be taken with the second specification. In other words, by insisting upon a theoretically distortion-free enevelope, some undesired sideband radiation is produced. However, this undesired radiation is relatively small and the wave is quite suitable for the applications indicated.

A SSB wave is a hybrid modulated wave. That is, SSB waves have both AM components and PM components. Thus, in order to describe an SSB wave, one must define the PM function as well as the AM function.

In the case of CSSB, we assume that the envelope is theoretically free of distortion. Thus, if a sine wave is fed to a perfect CSSB transmitter, the envelope of the radiated waves has a perfect sinusoidal shape. The other defining function of this hybrid modulated wave is the PM component, and research on the CSSB system centered upon the quest for an appropriate PM function. *What was desired was a PM wave that, when amplitude-modulated by the input signal, would concentrate the energy of the total wave in one sideband.* It will be seen that the PM wave is a nonlinear function of the input wave. The procedure used in deriving this PM wave will be described later.

In order to obtain the desired low distortion envelope function for large modulation percentages by a single tone, it is necessary to have at least three spectrum components. This, of course, is true for conventional AM where three components (carrier plus symmetrical upper and lower sidebands) are transmitted when a sinusoidal wave is used to modulate the transmitter. In the AM case, the carrier is located in the center of the spectrum and is surrounded by the upper and lower sideband components.

In the case of CSSB, the carrier is on one side of the wave and first- and second-order sideband components *are both on one side of the carrier.* These three components are necessary at high percentages of modulation but, as will be shown later, at low percentages of modulation only a carrier and a first-order sideband component are required. *Since the high-frequency components of speech and music have relatively small amplitudes, these high-frequency components require only one sideband plus the carrier, and the bandwidth of the CSSB signal can be made approximately equal to the bandwidth of the audio modulating signal.*

Since the procedure used in developing the CSSB system was relatively involved, it might be advantageous to describe briefly the steps taken in performing this work. First of all, analysis was performed on various three-element waves in order to determine the optimum component relationship of a three-element wave. *This*

three-element wave was used as a mathematical "model" for determining the structure of an actual CSSB wave for the case of single-tone modulation.

It might be questioned why the three-element "model" wave was not used as the actual wave. The reason is two-fold: First, such a wave, with the proper component phase and amplitude relationships, would be quite difficult to generate. Second, the three-element model has an inherent envelope distortion of almost 5 per cent. Because the derived CSSB wave is theoretically free of envelope distortion, and in practice has approximately only 1 per cent distortion, the derived system is actually an improvement on the mathematical model.

Once the desired relative amplitudes of the components were determined, it was necessary to calculate the PM component of this three-element mathematical model wave. After the PM component was derived, comparison was made between the PM characteristics of various *two-element carrier-plus-sideband waves* and the desired PM component. Two-element waves were considered because they are simple to produce. Study proved that the summation of the PM components of two different two-element waves closely approximated the desired three-element PM component.

A spectrum analysis was then performed showing that the synthetic PM component, when amplitude-modulated by a perfect sine wave function, would have satisfactory spectrum characteristics. At this point, experimental work was performed to ascertain the actual operating characteristics.

IV. Discussion of Three-Element CSSB Mathematical "Model" Wave and Means for Approximation

From the above discussion of a SSB full-carrier wave, we see that a two-component wave has an envelope function that would be highly distorted and unacceptable as a true compatible system. Therefore, it is necessary to transmit additional components in order to make the wave compatible with conventional AM receivers. In practice, one additional sideband component placed in the desired sideband and separated from the carrier by two times the frequency of a sinusoidal modulation is all that is necessary to make the wave compatible.

Fig. 5 shows the amount of envelope distortion at maximum percentage modulation of a three-element wave for various ratios of second-order sideband component. It is assumed that sinusoidal modulation is used and that the carrier level is varied so that the sum of the carrier plus the second-order sideband just equals the amplitude of the unmodulated carrier. A study of this curve indicates that the distortion varies from zero, at a ratio of 0.5, to about 24 per cent distortion at zero ratio of second-order to first-order sideband voltage. Of course, the 24 per cent envelope distortion figure occurs at the limiting situation where the signal degenerates into a two-element wave.

As mentioned above, the amplitude of the carrier frequency component in this three-element model wave equals the difference between unmodulated carrier amplitude and the second-order sideband component amplitude. Thus, in the 0.5 ratio, 100 per cent single-tone modulation case, the carrier level equals the amplitude of the second-order sideband, and except for a frequency shift, it is the same as a conventional 100 per cent modulated AM wave. The difference is that the carrier is at one side of the spectrum of the wave, instead of in the center as in conventional AM (see Fig. 6).

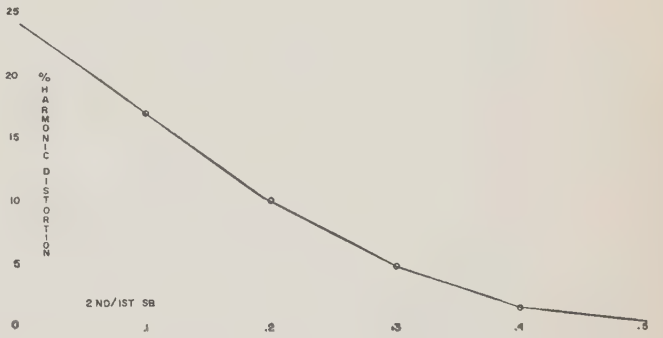


Fig. 5—Total harmonic distortion for three-element CSSB "model" wave as a function of the ratio of the second-order sideband to the first-order sideband at maximum modulation.

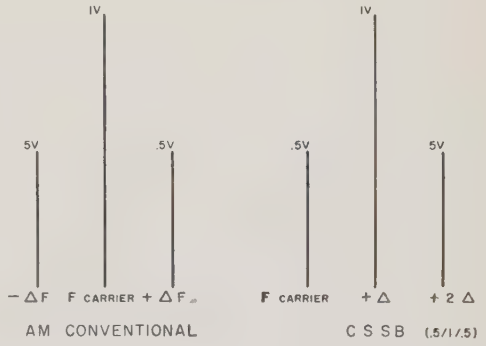


Fig. 6—Spectrum diagram of conventional AM wave and CSSB wave having 0.5 to 1 to 0.5 ratio of carrier to first-order sideband to second-order sideband.

Even though the 0.5-ratio model would appear to be the best choice because of the absence of envelope distortion, other requirements help to determine the choice. These other problems necessitate a lower ratio of second-order sideband to first-order sideband. If two audio signal tones of approximately equal amplitude are fed to the input of the CSSB transmitter, an undesired sideband radiation is produced. In this case, a lower ratio of second- to first-order sideband is preferred, and as will be shown later, a ratio of approximately 0.3 appears to be an acceptable compromise choice.

Another reason for using the carrier/first sideband/second sideband ratio of 0.7/1/0.3 is that at lower percentages of modulation the bandwidth of the signal must not increase. If a larger second-order component were used at maximum modulation, this would not be

true, and the undesired components would be excessive at lower percentages of modulation.

Thus, the three-element mathematical model of a CSSB wave may be described as follows:

$$e = E_{\text{carrier}}[(1 - 0.3m^2)\cos w_c t + m\cos(w_c + w_{Af})t + 0.3m^2\cos(w_c + 2w_{Af})t], \quad (1)$$

where

E_{carrier} = carrier voltage when modulation is absent,
m = envelope modulation factor,
w_c = angular velocity of carrier,
w_{Af} = angular velocity of tone modulation.

At low percentages of modulation, (1) indicates that the second-order component is very small. This can be seen by studying Figs. 1 and 2, which indicate that at low percentage modulation the envelope distortion of a conventional carrier plus a sideband two-element wave is small. Therefore, the requirement of a second-order term decreases as the percentage modulation decreases (see Fig. 7).

Eq. (1) shows that the first-order sideband is a linear function of percentage modulation and that the second-order sideband follows a squared function of percentage modulation. The carrier amplitude at low percentage modulation equals the average amplitude of the wave and gradually decreases as the modulation level increases. Though the model only provides approximate relationships, it appears to fit practical situations quite well, and it has been helpful in deriving some of the techniques used.

It should be stressed that the average amplitude of the actual CSSB wave is constant and does not shift with percentage modulation. In other words, no "carrier shift" is created when CSSB is used. Of course, if the transmitter produces distortion with attendant carrier shift when transmitting conventional amplitude signals, it will produce the same carrier shift when the CSSB system is used.

It should be further stressed that in the actual CSSB system, the mechanism for producing the envelope function is such that theoretically the system is completely free of envelope distortion, and is in that respect superior to the mathematical three-element model. Therefore, assuming a receiver having perfect characteristics, the entire system would be completely free of any harmonic or intermodulation distortion. In practice, measurements at a number of stations show that the total harmonic envelope distortion of the transmitter plus CSSB adapter is within 1 per cent of the distortion of the transmitter.

We have shown that at low percentage modulations only a carrier plus SSB is transmitted. Since the higher frequency components of voice and musical waves are generally of very low amplitude, the spectrum requirement for such an SSB wave is approximately equal to the audio bandwidth, and a wave closely approximating the bandwidth of conventional SSB is generated. When low-frequency high-amplitude signals are transmitted, the second-order sideband is required. But because this second-order sideband falls well within the bandwidth of the signal, it does not increase the bandwidth of the signal but merely increases the density of the spectrum of the wave.

Fig. 7 shows the manner in which the first-order and second-order sidebands vary with percentage modulation, assuming the model defined in (1). We note that, below 30 per cent modulation, the second-order sideband is more than 30 db below the carrier level. This is the reason the spectrum is maintained essentially equal to the audio bandwidth of the modulation.

Fig. 8 was derived from Fig. 7 by utilizing the frequency characteristic of music as illustrated in Fig. 4. The frequency axis indicates the frequency of the component being considered. Thus, at 5 kc one sees that the first-order sideband is 15 db below the amplitude of the low-frequency reference. A 2.5-kc signal produces the second-order component which would appear at 5 kc, and which is 25 db below the low-frequency reference and 10 db below the 5-kc first-order sideband. In addition, the effect of the sideband filter used in this system is shown cutting off the wave sharply above 8 kc. Fig. 8 shows that the first-order sideband, except for the very edge, is considerably greater than the second-order sideband. *Therefore, the second-order sideband does not materially alter the bandwidth of the signal because the band is already occupied by the first-order sideband.* This con-

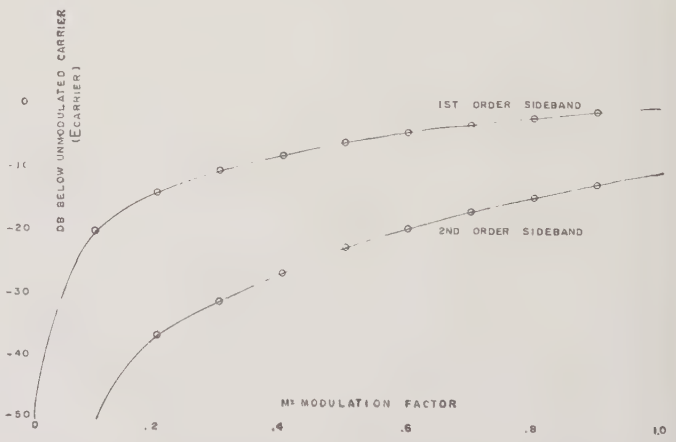


Fig. 7—Curve of amplitude of first-order and second-order sideband components for various amounts of modulation assuming "model" wave having carrier to first-order sideband to second-order sideband ratio of 0.7 to 1 to 0.3 at maximum modulation.

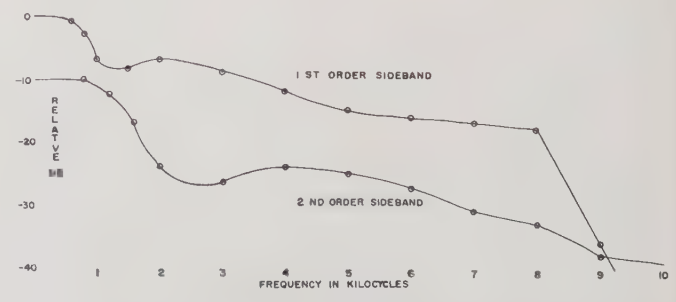


Fig. 8—Amplitude of first-order and second-order sideband components of "model" CSSB wave, assuming spectrum distribution of music shown in Fig. 4.

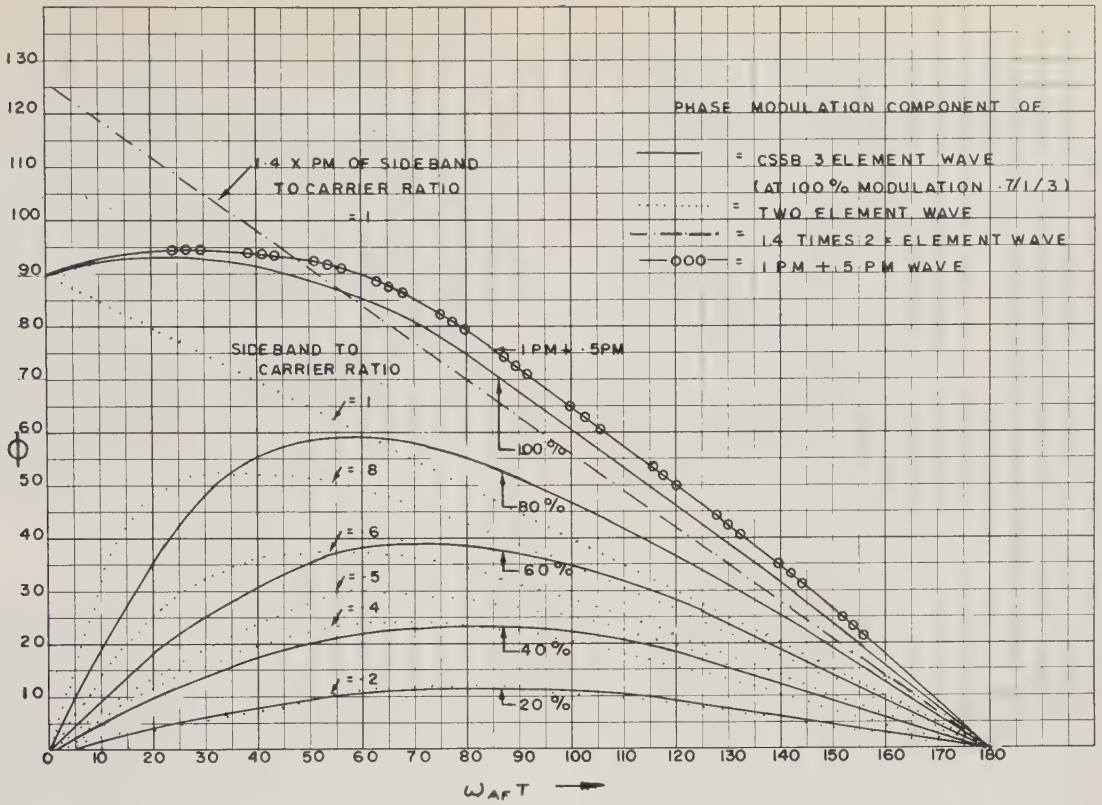


Fig. 9—Phase modulation component of CSSB "model" wave, also of two-element carrier plus sideband wave for various percentage modulation.

clusion is supported by extensive measurements which prove that the spectrum is merely made denser by the second-order sideband and does not increase the bandwidth appreciably over what a normal SSB system would require. It should be recalled that the fact that high-frequency music components are weak is also the basis for the use of pre-emphasis in FM broadcasting and disk recording.

Fig. 9 shows, among other things, a plot of the PM component of the three-element tone model wave having a carrier-to-first-order sideband-to-second-order sideband ratio of 0.7/1/0.3. It should be noted that in Fig. 9 a half cycle of the PM component is shown. This wave is an odd function about $t=0$, that is, $f(t)=-f(-t)$.

Fig. 9 also shows curves of the PM component of the model signal at low percentage modulation. Eq. (1) was used in calculating these curves.

The dotted curves shown in Fig. 9 represent the PM component of a two-element wave having a carrier-to-sideband ratio as indicated on the curve. It is seen that, except at very low percentages of modulation, the PM for the two-element case is considerably different from that for the desired CSSB case.

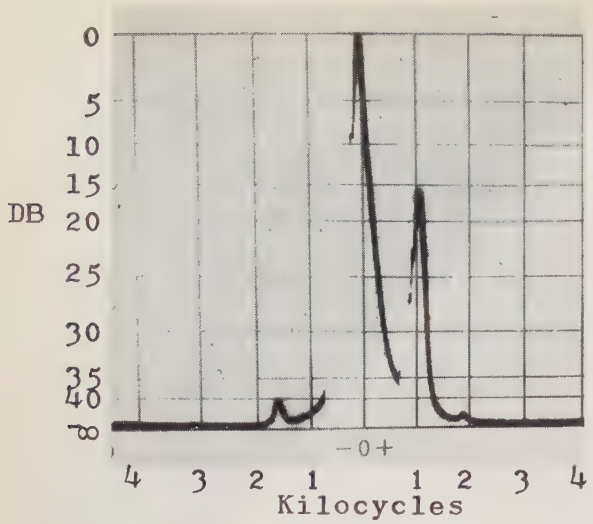
The phase-modulated wave produced by adding the PM component of a two-element wave having a carrier-to-sideband ratio of 1 to the PM component of a two-element wave having a carrier-to-sideband ratio of 0.5 closely approximates the PM component of the desired CSSB wave at 100 per cent modulation. At moderate percentages of modulation, the PM component of a two-element wave has too large an amount of phase swing for certain portions of the modulating cycle, and too small a phase swing at other segments of the modulating cycle. In order to satisfy this problem, nonlinear circuits were designed to produce a summation of two PM components, which closely approximate the PM component of the two- and three-element CSSB models.

For instance, at 60 per cent modulation, a sideband-to-carrier ratio of 0.5 would be added to one of about 0.1. Thus, by proper use of nonlinear curves a close fit can be obtained for various percentages of modulation. The panoramic pictures shown in Fig. 10 indicate that the technique used was successful in producing the proper PM component. Thus, this PM wave when amplitude modulated by a wave derived from the input wave produces the desired spectrum characteristic.

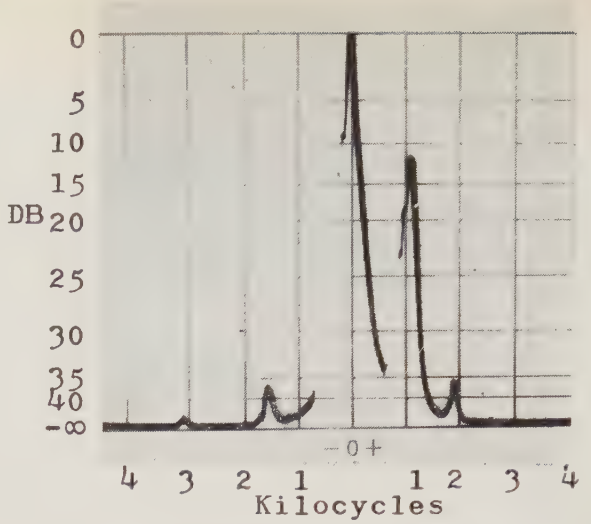
Fig. 10(a)–(o) (pp. 1510–1512) were made by taking panoramic pictures of a small laboratory CSSB transmitter. In taking the pictures in Fig. 10(a)–(l), this transmitter was modulated at various percentages of modulation with a 1000-cycle tone. The pictures indicate that the undesired sideband is approximately 33 db below the ummodulated carrier under the worst operational condition. Note that Fig. 20 (k) and (l) show the spectrum when the transmitter is over-modulated.

Fig. 10(m) shows the same laboratory transmitter 100 per cent modulated by white, or random, noise. Fig. 10(n) shows the spectrum for 100 per cent modulation of this transmitter by program material and Fig. 10(o) shows the effect of over-modulation; that is, on peaks the transmitter indicated 110 per cent modulation.

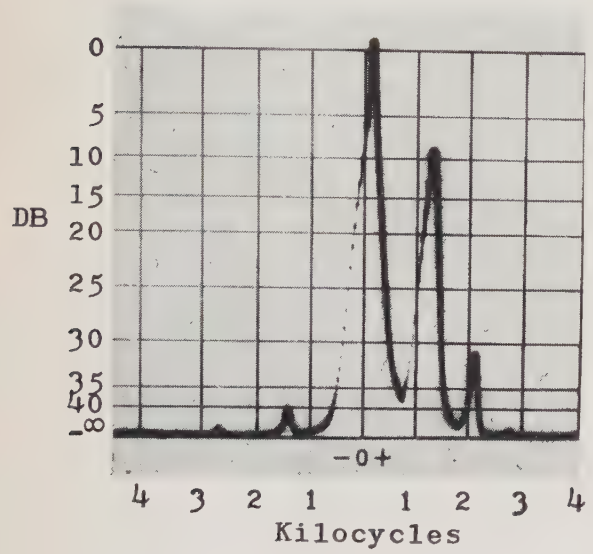
The problem of generating such a CSSB wave re-



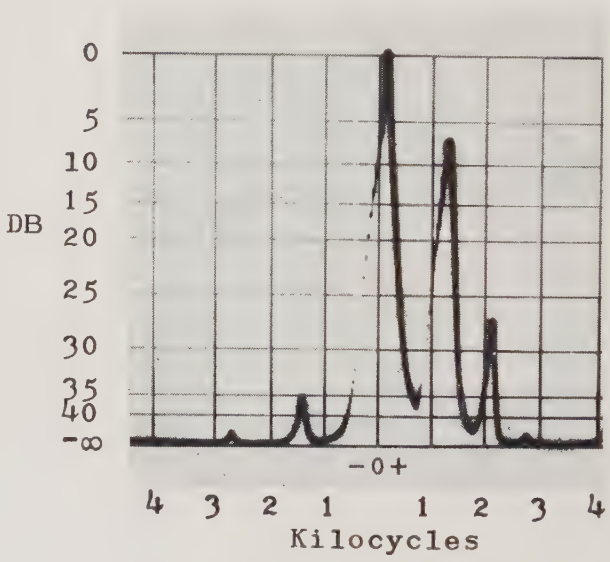
(a) Single tone, 10 per cent modulation



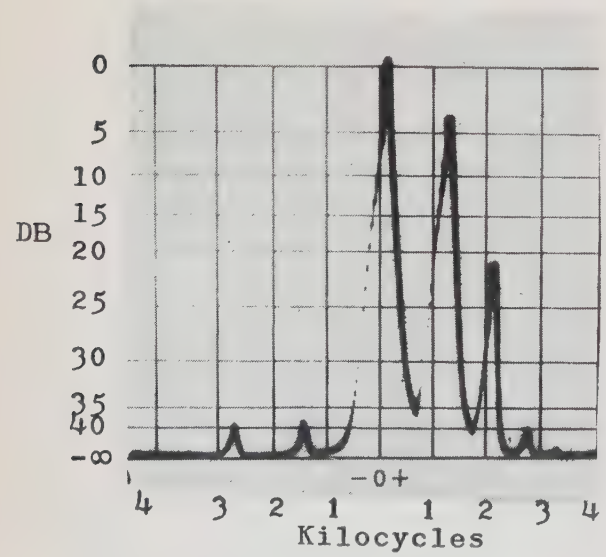
(b) Single tone, 20 per cent modulation



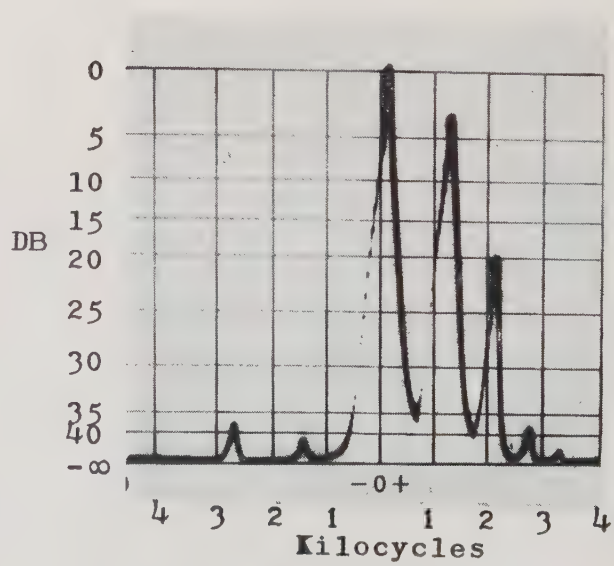
(c) Single tone, 30 per cent modulation



(d) Single tone, 40 per cent modulation

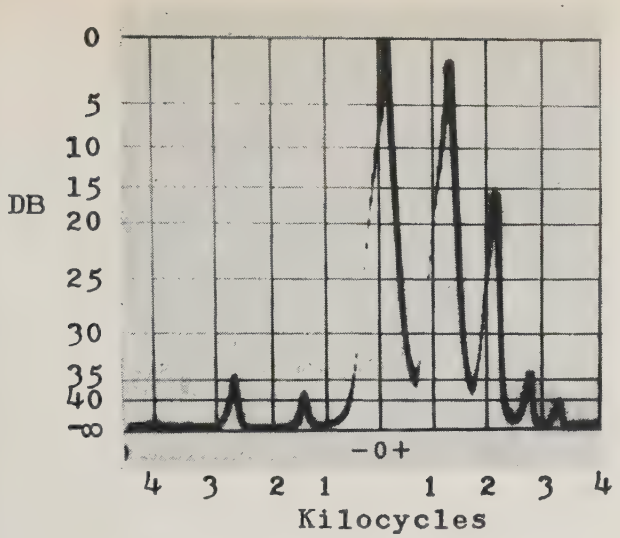


(e) Single tone, 55 per cent modulation

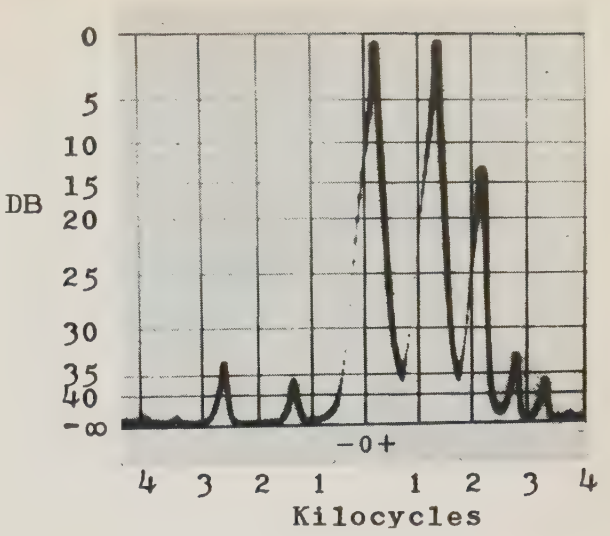


(f) Single tone, 60 per cent modulation

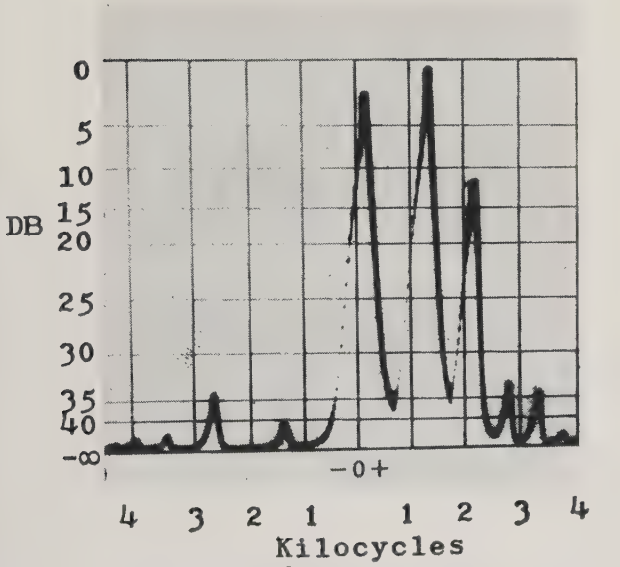
Fig. 10—Panoramic photographs of actual CSSB wave for various modulation conditions.



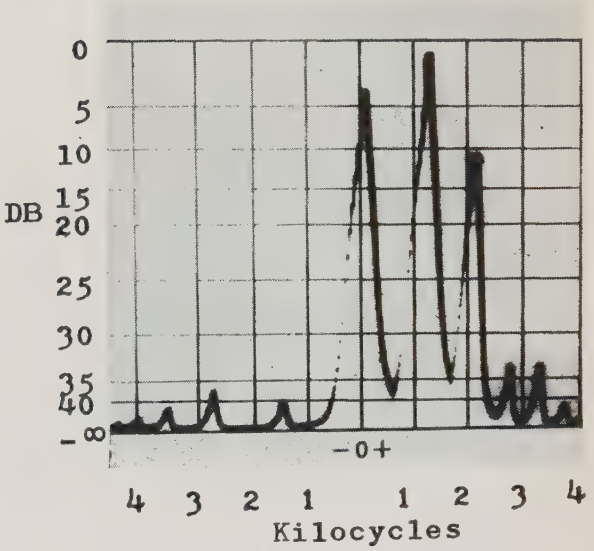
(g) Single tone, 70 per cent modulation



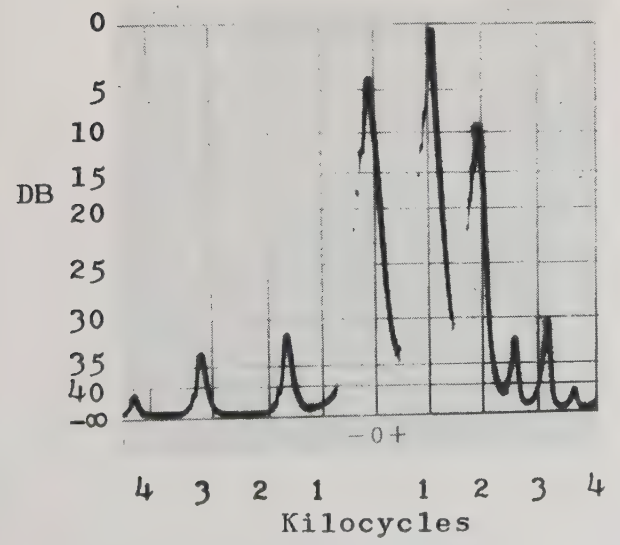
(h) Single tone, 80 per cent modulation



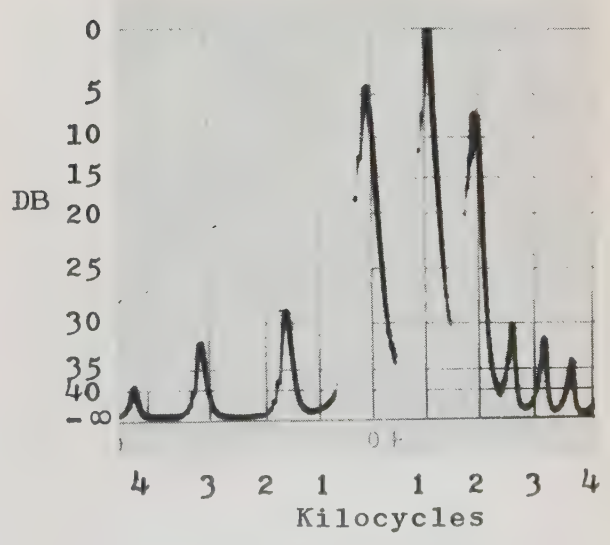
(i) Single tone, 90 per cent modulation



(j) Single tone, 100 per cent modulation

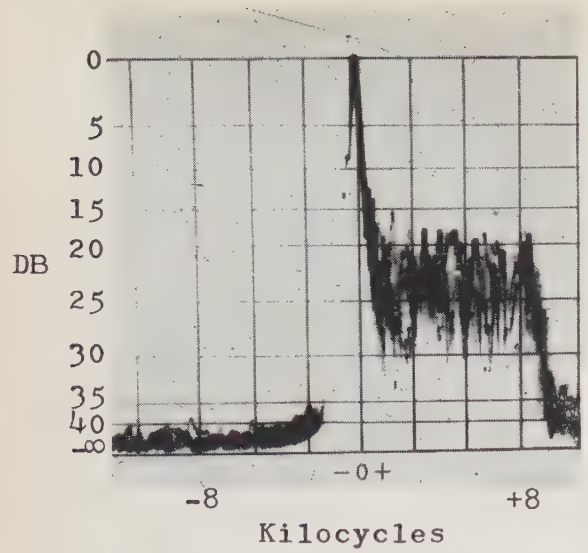


(k) Single tone, 110 per cent modulation

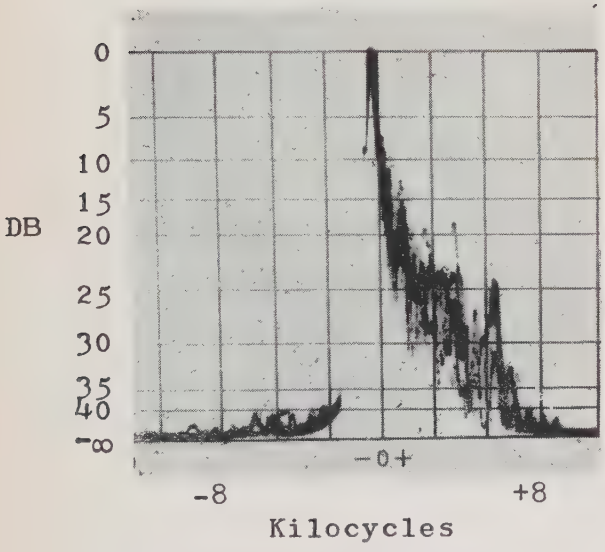


(l) Single tone, 120 per cent modulation

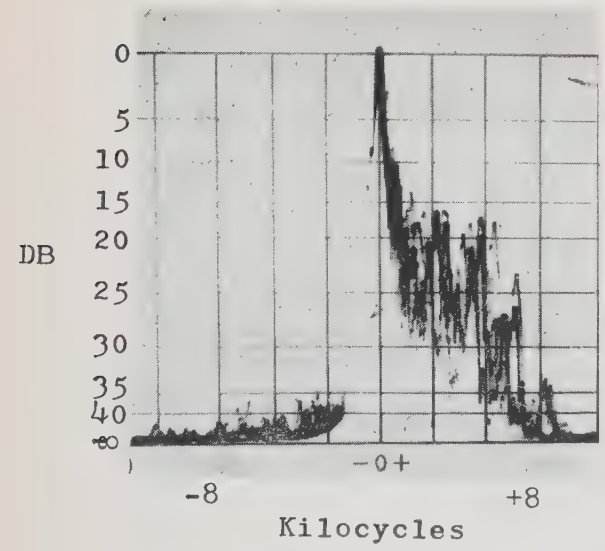
Fig. 10—(Continued)



(m) 100 per cent white noise modulation



(n) 100 per cent program modulation



(o) 110 per cent program modulation

Fig. 10

solves itself into producing the desired PM. This can be done by a number of methods.

A CSSB technique (Fig. 11) which was extensively used in equipment produced by the Kahn Research Laboratories was to generate a normal SSB full carrier wave and pass this wave through a limiter in order to isolate the PM. The amount of PM can be shown to be insufficient by study of Fig. 5 from a previously published article[4] on the envelope elimination and restoration SSB system. (Also, see Fig. 9 of this paper.) Therefore, the amount of PM was multiplied by use of a harmonic multiplier and a frequency divider. The combination increased the PM by a factor of 1.4. Though this technique produced acceptable signals, it had one severe problem. It was very sensitive to overmodulation because the average frequency suddenly shifts when the sideband exceeds the amplitude of the carrier at the output of the SSB generator.

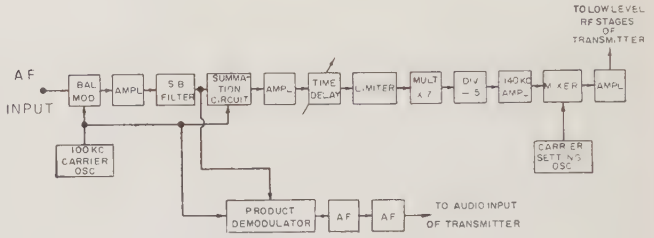


Fig. 11—Block diagram of CSSB system using 1.4 phase stretch circuit.

A more recent development resulted in the "1 PM plus 0.5 PM" system which takes considerable amounts of overmodulation. This system produces a wave having reasonably good sideband isolation during overmodulation periods and excellent sideband isolation during normal operation. This system is described in detail in Section V.

A special technique for communication purposes, which has some inherent envelope distortion but very low out-of-band radiation, has also been investigated but will not be discussed here.

V. Block Diagram of CSSB System Using 1 PM Plus 0.5 PM Technique

The block, shown in Fig. 12, indicates the procedure used to produce the desired CSSB wave by the new 1 PM plus 0.5 PM procedure. The SSB wave is generated in a conventional SSB generator utilizing a balanced modulator plus sideband filter. In the actual equipment, a high-selectivity crystal lattice network filter is utilized. The amplitude characteristic of this filter is such that the output is essentially flat from 50 cycles to 8000 cycles above the carrier frequency. The SSB wave, without carrier, is amplified and fed to a cathode follower. This cathode follower provides a low output impedance and is used to drive two nonlinear circuits which have similar, but not equal, nonlinear characteristics. Both nonlinear networks provide lower attenuation for high-amplitude wave levels than they do for

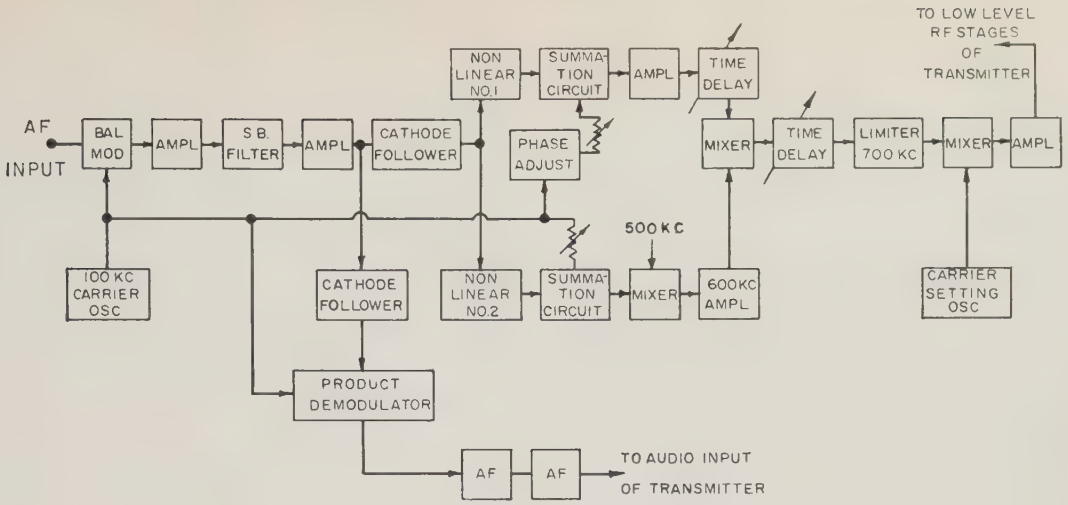


Fig. 12—Block diagram of CSSB system using 1 PM+0.5 PM technique.

low-amplitude waves. These networks are required because the amount of PM boost varies from about 1.5 at 100 per cent modulation to unity at very low percentages of modulation.

Nonlinear circuit No. 1 feeds a summation circuit in which the carrier is added to the sideband wave. The carrier is passed through a phase-shift circuit to align properly the phase relationship between the two paths. The carrier amplitude is such that when the input level is just sufficient to produce 100 per cent modulation, the carrier and the sideband are equal. This is required to produce the PM wave in Fig. 9 which is labeled "sideband-to-carrier ratio equals 1." The output of this summation circuit is amplified and fed through a time delay circuit. The time delay circuit is used to compensate for small difference in time delay between the 1 PM path and the 0.5 PM path. The output of the time delay circuit is fed to a mixer.

Now returning to the output of the cathode follower, it is seen that the cathode follower also drives nonlinear circuit No. 2. This nonlinear circuit feeds a second summation circuit in which the carrier at a level of twice the sideband level at maximum modulation is added to the sideband wave. Thus, the sideband-to-carrier ratio at 100 per cent modulation is 0.5. The output of the summation circuit feeds a mixer in which the wave is heterodyned, in this example, to 600 kc. The output of the 600-kc amplifier is fed to the mixer, in which it heterodynes with the 100-kc 1 PM wave and the mixer summation component 700 kc is selected.

The 700-kc component is phase-modulated by the sum of the PM of the two paths so that the PM equals 1 PM plus 0.5 PM at 100 per cent modulation. At very low percentages of modulation, the nonlinear circuits function so as to produce a PM component equal to that of carrier plus sideband wave, where the sideband-to-carrier ratio equals the percentage modulation. This PM is proper for minimizing the unwanted sideband, and it closely approximates the PM waveshapes of the three-element wave shown in Fig. 9. This combined AM and PM wave is then fed to a limiter in which the AM component of the wave is removed, producing a pure PM wave. The wave is then converted in frequency to a desired carrier frequency where it may be used to drive a class-C high-level, low-level, Doherty, Chireux, or Ampliphase type of transmitter.

In the transmitter, the 1 PM plus 0.5 PM wave is amplitude-modulated by a wave form that is derived from the SSB wave by demodulating that wave by the carrier in a product demodulator. Since a product demodulator may be used to produce a very low distortion audio wave when demodulating a conventional SSB wave, the envelope can be almost free of envelope nonlinear distortion. The reason for this rather complicated technique for deriving the AM component is that, while the original audio wave has all the desired spectrum components, the phase of these components is altered by the crystal filter in the SSB generator. Therefore, unless an elaborate compensating phase-shift network is used, such a wave would not have the proper phase characteristics for the desired envelope function.

Because the CSSB system described here is synthesized from PM and AM components, any type of AM transmitter now in current use may be adapted for CSSB use with little or no alteration of the transmitter.

VI. Spectrum Analysis of a CSSB Wave

Those familiar with modulation analysis techniques will recognize that a complete analysis of the spectrum characteristic of such a system is an extremely involved problem. The reasons for this are that the block diagram incorporates a number of nonlinear elements, and that PM is used with its infinite spectrum. Therefore, in the following analysis only the relatively simple case of single-tone modulation is discussed. *This analysis does not pretend to be complete, but is shown merely as an aid in understanding the system whereby the wave is generated* and because it was used in the original development of the system. Further analysis with different types of waveshapes would no doubt produce interesting results;

however, it is expected that such work would be quite difficult and laborious. Furthermore, the researcher must still face the problem of determining what is a "typical" waveshape, that is, one which will simulate voice and music. Because of the large number of voice and music waveshapes involved, such solutions would always be subject to question.

A most interesting technique[14] was utilized by Mertens of the European Broadcasting Union in analyzing CSSB. He used an analog computer and generated a large number of operating situations. Because full information had not been published on the CSSB system, Mertens recognized that the analysis was limited. It should be noted that the use of an analog computer on the full detailed block of this system would require considerable complexity.

The following describes the procedure used in determining the CSSB spectrum for a single tone modulating wave. Since the worst spurious radiation would generally occur at the maximum modulation, this is the condition assumed in the following analysis.

The spectrum of the amplitude-limited sum of two equal-amplitude sinusoids may be represented as follows:[5]

$$e_{\text{pm}1/1} = \sum_{n=-\infty}^{n=\infty} A_n \cos (w_{c1} + nw_{Af})t \tag{2}$$

where w_{c1} is the angular velocity of the carrier frequency and w_{Af} is the angular velocity of the modulating audio tone.

$$\begin{aligned} A_0 &= +1 & A_{+1} &= -1/3 \\ A_{-1} &= +1 & A_{+2} &= +1/5 \\ A_{-2} &= -1/3 & A_{+3} &= -1/7 \\ A_{-3} &= +1/5 & A_{+4} &= +1/9 \\ A_{-4} &= -1/7 & & \text{etc.} \\ & \text{etc.} \end{aligned}$$

The amplitude of the components has been normalized so that the carrier-frequency component A_0 and the first-order lower-sideband component A_{-1} have unity amplitudes. It should be noted that it is assumed that the lower sideband was the desired sideband in the present case, although either sideband may have been considered in the analysis. Also, the instant of time chosen for $t=0$ is when the carrier and the lower sideband phasors are exactly in phase.

It is interesting to note that this spectrum is equivalent to that of a double-sideband suppressed-carrier signal in which the modulating signal is a square wave. The fundamental frequency component of the square wave would equal $w_{Af}/2$.

The spectrum of the PM component of an SSB wave having a sideband-to-carrier ratio of 0.5 may be represented as

$$e_{pm}1/0.5 = \sum_{m=-\infty}^{m=\infty} B_m \cos (w_{c2} + mw_{Af})t. \tag{3}$$

The value of B_m may be found in some very detailed tables calculated by Granlund[15] or by utilizing a technique described in a paper analyzing limiters.[16] The tuned circuits, in the plate circuit of the mixer, are tuned to a frequency which is the sum of the two input frequencies. Therefore, a wave is produced having a PM component equal to the sum of the PM component of the 1-to-1 sideband-to-carrier ratio signal plus the PM component of the 0.5-to-1 sideband-to-carrier ratio.

$$\begin{aligned} & e_{PM}1/1 \times e_{PM}1/0.5 \\ &= \sum_{n=-\infty}^{n=\infty} A_n \cos (w_{c1} + nw_{Af})t \times \sum_{m=-\infty}^{m=\infty} B_m \cos (w_{c2} + mw_{Af})t \\ &= \sum_{n=-\infty}^{n=\infty} \sum_{m=-\infty}^{m=\infty} A_nB_m(\cos w_{c1} + nw_{Af})t \times \cos (w_{c2} + mw_{Af})t \\ &= \sum_{n=-\infty}^{n=\infty} \sum_{m=-\infty}^{m=\infty} \frac{A_nB_m}{2} \cos [w_{c1} + w_{c2} + (n+m)w_{Af}]t \\ &\quad + \frac{A_nB_m}{2} \cos [w_{c1} - w_{c2} + (n-m)w_{Af}]t. \end{aligned} \tag{4}$$

Since the mixer output is filtered, only those components that fall close to the summation frequency of the two input frequencies, that is, $w_{c1}+w_{c2}$, are passed. Therefore, the second term of (4) may be ignored.

The chart in Fig. 13 indicates the procedure used in calculating the resultant PM spectrum. The top row indicates the PM spectrum of the two equal tone waves (1 PM), and the first left-hand column indicates the PM spectrum of the carrier-plus-half-amplitude-lower-sideband wave. The product of these spectrum components is shown in the center of the chart. (These figures were not multiplied by 0.5 as called for by the trigonometric identity, since we are only interested in relative values.) Each of these numbers contributes to its columns a final amplitude which is shown in the bottom row with the amplitude of the new frequency components. This wave centers around the sum frequency of the first carrier frequency and the second carrier frequency ($w_{c1}+w_{c2}$). For instance, if the third-order lower sideband were multiplied by the first-order upper sideband, the resulting frequency would be equal to the sum of the carrier frequencies minus twice the audio frequency. The products with appropriate signs are then summed and the resulting summation is the spectrum composition of the total PM wave. This is the PM spectrum which is amplitude-modulated in the associated transmitter.

Fig. 14 shows the effect of AM on the PM wave. The result of this calculation provides us with the spectrum

[14] H. Mertens, "A study by means of an analogue computer of the spectrum of CSSB modulation," *Eur. Broadcasting U. Rev.*, vol. 64, pp. 249–258; December, 1960.

[15] J. Granlund, "Interference in Frequency Modulation Reception," Research Lab. of Electronics, M.I.T., Cambridge, Mass., Tech. Rept. 42, pp. 29–48; January 20, 1949.

[16] L. R. Kahn, "Analysis of a limiter as a variable gain device," *Elec. Engr.*, vol. 72, pp. 1106–1109; December, 1953.

PM Component of Carrier plus equal SB wave	$\omega_{c1}-4\omega_{Af}$ −0.143 v	$\omega_{c1}-3\omega_{Af}$ +0.200	$\omega_{c1}-2\omega_{Af}$ −0.333	$\omega_{c1}-\omega_{Af}$ +1.0	ω_{c1} +1.0	$\omega_{c1}+\omega_{Af}$ −0.333	$\omega_{c1}+2\omega_{Af}$ +0.200	$\omega_{c1}+3\omega_{Af}$ −0.143	$\omega_{c1}+4\omega_{Af}$ +0.111
PM Spectrum of Carrier Plus Half Amplitude SB Wave									
$\omega_{c2}+6\omega_{Af}+0.0031$								+001	−001
$\omega_{c2}+5\omega_{Af}-0.0068$				+001	−001	+001	−001	+002	−007
$\omega_{c2}+4\omega_{Af}+0.0150$			−001	+002	−002	+003	−005	+015	+015
$\omega_{c2}+3\omega_{Af}-0.0345$		+003	−004	+005	−007	+011	−035	−035	+011
$\omega_{c2}+2\omega_{Af}+0.0834$	−008	+009	−012	+017	−028	+083	+083	−028	+017
$\omega_{c2}+\omega_{Af}-0.2252$	−025	+032	−045	+075	−225	−225	+075	−045	+032
$\omega_{c2}+0.9342$	−133	+187	−311	+934	+934	−311	+187	−133	+104
$\omega_{c2}-\omega_{Af}+0.2586$	+052	−086	+259	+259	−086	+052	−037	+029	−024
$-2\omega_{Af}-0.0335$	+011	−034	−034	+011	−007	+005	−004	+003	
$-3\omega_{Af}+0.0085$	+009	+009	−003	+002	−001	+001	−001		
$-4\omega_{Af}-0.0027$	−003	+001	−001						
	−097	+121	−152	+1.306	+577	−380	+262	−191	+147
	$-4\omega_{Af}$	$-3\omega_{Af}$	$-2\omega_{Af}$	$-\omega_{Af}$	$\omega_{c1}+\omega_{c2}$	$+\omega_{Af}$	$+2\omega_{Af}$	$+3\omega_{Af}$	

Fig. 13—Spectrum calculations for determining PM spectrum of 1 PM+0.5 PM system.

PM Component of 1 PM+0.5 PM	$-4\omega_{Af}$ −097	$-3\omega_{Af}$ +121	$-2\omega_{Af}$ −152	$-\omega_{Af}$ +1.306	$\omega_{c1}+\omega_{c2}$ +577	$+\omega_{Af}$ −380	$+2\omega_{Af}$ +262	$+3\omega_{Af}$ −191	$+4\omega_{Af}$ +147
Sideband Produced by Modulation of each PM Component by Sine Wave at 100 per cent Modulation	+038 +061	−049 −076	+061 +653	−0.076 +0.289	+653 −190	+289 +131	−190 −096	+131 +074	−096 −069
Spectrum Produced	+002	−004	+562	+1.519	+1.040	−040	−024	+014	−018
Spectrum Produced in db	−57.7	−51	−8.6	0	−3.3	−31.7	−36	−40.7	−38.6
			Second-Order Sideband	First-Order Sideband	Carrier				

Fig. 14—Spectrum calculations for CSSB system assuming perfect transmitter.

PM Component of 1 PM+0.5 PM	+075	$-4\omega_{Af}$ −097	$-3\omega_{Af}$ +121	$-2\omega_{Af}$ −152	$-\omega_{Af}$ +1.306	$\omega_{c1}+\omega_{c2}$ +577	$+\omega_{Af}$ −380	$+2\omega_{Af}$ +262	$+3\omega_{Af}$ −191	$+4\omega_{Af}$ +147
Sideband Produced by Sinusoidal 97 per cent AM of Each PM Component	−023 −047	+036 +059	−047 −074	+059 +633	−074 +280	+633 −184	+280 +127	−184 −093	+127 +071	−093
Sidebands Produced by 3 per cent AM Second Harmonic Distortion Wave of Each PM Component		+001 +002	−001 −020	+002 −008	−002 +006	+002 −004	−020 +003	−009 −002	+006 +002	−004
Spectrum Produced		+001	−021	+534	+1.516	+1.024	+010	−026	+015	
Spectrum Produced in db		−63	−37.1	−9	0	−3.4	−43.6	−35.4	−40.1	
				Second-Order Sideband	First-Order Sideband	Carrier				

Fig. 15—Spectrum calculations for CSSB system assuming transmitter having 3 per cent envelope distortion.

composition of the desired CSSB wave. Each of the PM spectrum components is assumed to be a carrier for the AM; therefore, two symmetrical sidebands spaced by the frequency of modulation are produced about each PM component. The summation of the sidebands produced by the AM and the PM spectrum is the spectrum of the desired-output wave. It should be noted that we have assumed, in this figure, that there was no envelope distortion present. Fig. 15 provides a similar analysis except for the assumption that the transmitter has 3 per cent second harmonic[17] envelope distortion.

We have assumed, in calculating the PM spectrum, that separate limiters were used in the 1 PM branch and

[17] The distortion figure of 3 per cent is based on the economic factors involved in high-powered transmitter design. Economic use of large tubes is necessary and therefore a small degree of emission limiting is generally experienced. This is the basis for the assumption that the second harmonic distortion is out of phase with the fundamental component during the positive modulation peak.

the 0.5 PM branch. Experiments were performed with this type of block diagram which proved that the same spectrum results whether the limiting is performed on the individual components or on the heterodyned wave. Of course, the analysis could have been performed by merely mixing the two waves and then determining the effect of limiting after the heterodyning process by use of a technique outlined in a previous paper.[16]

Another undesired sideband component is produced when more than a single tone is transmitted. The following analysis indicates the level of this component.

The level of the carrier component is a function of the audio input level. If two equal-amplitude tones, which have a combined peak level sufficient to cause 100 per cent modulation, are used to modulate the transmitter, the percentage of modulation will vary from 0 per cent to 100 per cent according to the phase relationship between the two modulation tones. When the two components are instantaneously out of phase their resultant is zero, and the instantaneous percentage modulation is zero per cent. At other times the two components add and the modulation reaches 100 per cent. Thus, the carrier amplitude (for the 0.7/1/0.3 carrier-to-sideband component ratio) shifts from 0.7 at 100 per cent modulation to 1 at zero per cent modulation. This variation in carrier level occurs at a rate equal to the beat frequency between the two modulation tones.

Of course, the variation in carrier amplitude is AM, and therefore two sidebands are generated. These components are each equal to $(1-0.7)/4$ E_{carrier} or 0.075 E_{carrier}. Thus, each of these components is a little over 23 db below the unmodulated carrier amplitude or 29 db below the peak envelope power level. One of these sidebands falls in the undesired sideband spectrum. As the two-tone audio wave decreases in amplitude, the undesired sideband component decreases. Also, as the two tones depart from amplitude equality the component decreases.

In practice, the spacing between two strong equal amplitude tones in speech and music normally is very small,[18] less than 100 or 200 cycles, and therefore the spurious components produced by this mechanism are very close to the carrier and should not increase the adjacent-channel interference to any great degree.

It should be stressed that while this spurious component is undesirable, it does not introduce envelope distortion and is not heard or measured when such a signal is received on an envelope detector. Actually, all of these spurious components are required if the envelope is to be free of distortion, and if the receiver eliminates these components some distortion will result. Fortunately, those components are extremely close to the carrier and, in practice, are passed by the receiver so that the envelope is relatively free of distortion.

[18] Harvey Fletcher, "Speech and Hearing," D. Van Nostrand Co., Inc., Princeton, N. J., 2nd ed.; 1953.

ADVANTAGES AND EXPERIMENTAL EVALUATION OF THE CSSB SYSTEM

One of the main applications of CSSB is in the field of AM broadcasting. Therefore, detailed measurements under broadcasting conditions are given below.

All the tests made were based on the original 1.4 phase stretch circuitry. From subsequent tests, it would appear that use of the new CSSB technique, 1 PM plus 0.5 PM, would provide additional advantages.

VII. "Shaped Noise" Evaluation Procedure

One of the basic problems in attempting a quantitative comparison of equipment developed for use in the transmission of voice and music is determination of a method for evaluating the effect of the characteristic spectrum distribution of the signal. Obviously, the use of a constant amplitude sine wave up to 15 kc would be unrealistic and would preclude acceptance of our present interference standards which allow 10-kc spacing of carriers. Therefore, it was felt that a new technique was required for proper comparison of CSSB with AM. The use of this technique has been favorably commented upon by one distinguished engineer,[19] and it may be useful for evaluation of other program-modulated signals.

This new technique might be called a "shaped noise" evaluation procedure. In carrying out this procedure, a filter was designed and built which has the musical energy frequency characteristic shown in Fig. 4. This filter was then fed white noise and the output was used in the evaluation tests. By this procedure, the statistical measurement problem was bypassed and one answer was accepted for this spectrum characteristic. Of course, the accuracy of the evaluation is largely dependent upon the shape of the curve used for shaping the noise. It should be mentioned that there is a heated controversy raging over the exact nature of the spectrum energy distribution of voice and music. Since our own tests are quite limited, we are forced to depend upon the authority of others.

It might be mentioned, furthermore, that one recent publication[20] questions the possibility of ever deriving a meaningful spectrum distribution curve. Thus, we see that this is not a simple matter; but neither is an evaluation simple which is made up of a large number of samples which may in themselves prove in the long run to be atypical. A study of various published spectrum diagrams relative to the peak values of *the total audio waves* indicates that the probability is extremely small that a component above 3 or 4 kc would have sufficient amplitude to produce a strong second-order sideband component (see Fig. 4).

[19] R. N. Harmon, "Reply Comments of Westinghouse Broadcasting Co., Inc.," FCC Docket No. 13596 (RM-156); January 13, 1961.

[20] J. G. McKnight, "The distribution of peak energy in recorded music, and its relation to magnetic recording systems," *J. Audio Engrg. Soc.*, vol. 7; April, 1959.

VIII. Signal-to-Noise Ratio and Frequency Response Comparison

In order to determine the amount of improvement in high-frequency response and signal-to-noise ratio of the CSSB system, consideration will be given first to the use of receivers having square top-selectivity characteristics. Later, measurements of conventional inexpensive broadcast receiver characteristics will be considered.

If a *perfect* AM receiver were used to receive a double-sideband AM wave, its pass band would have to be two times the highest audio frequency to be received. Thus, if an AM wave is modulated by components up to 10 kc, the IF and RF stages of the receiver should be 20 kc wide. For CSSB transmission, the bandwidth need only be 10 kc wide. Therefore, since random noise power is doubled when the bandwidth is doubled, the signal-to-noise ratio would be 3 db better for CSSB transmission. In the case of impulse noise, the noise power increases as a squared function of bandwidth. Therefore, there is a 6 db S/N ratio improvement for impulse noise when CSSB transmission is used.

It should be noted that these improvements in signal-to-noise ratios are for AM receivers with square top-selectivity characteristics, and the receivers in use today are far from perfect in regard to their selectivity characteristics. Actually, in practice these receivers are much narrower than they should be for double-sideband use for economic reasons; they therefore tend to destroy signal components that are important to speech intelligibility.

It becomes very difficult to specify signal-to-noise improvements when one considers the relative importance of different voice and music spectrum components. This is especially true in situations where the selectivity characteristic of the networks involved is rounded.

For instance, if a voice component at 3500 cycles is attenuated 7 db and a component at 2000 cycles is attenuated 1 db, then the signal-to-noise ratio for that 3500-cycle component would be reduced over the 2000-cycle tone by a factor of 6 db. Thus, the effective signal-to-noise ratio is a function of frequency response, and a single figure for signal-to-noise ratio cannot be easily derived. Therefore, no exact figure will be offered for the signal-to-noise improvement of CSSB, because the signal-to-noise ratio and the frequency response of the system are so closely related.

Except for the difference in bandwidth, the frequency response of the square selectivity curve AM receiver would be identical to that of a square selectivity curve CSSB receiver. Therefore, there would be no frequency response difference between the two systems when such receivers were used.

As noted above, the bandwidths of both domestic sets and foreign sets leave much to be desired. Actually, their poor high-frequency characteristics not only degrade the musical reproduction process but also reduce voice intelligibility. It has been pointed out[18] that loss of frequency response above 3000 cycles reduces the intelligibility of "s" sounds from 100 per cent to 40 per cent, the "th" sound from 100 per cent to 66 per cent, the "z" sound from 100 per cent to 80 per cent, and the "t" sound from 100 per cent to 81 per cent. These measurements were made under relatively good signal-to-noise conditions; it has been noted that under poor signal-to-noise conditions the loss of intelligibility due to reduced high-frequency response is even more severe. There is a noise-masking effect that further degrades intelligibility when high-frequency components are attenuated by the limited frequency response of the equipment.

The CSSB system, by improving the high-frequency response of the system, increases the area of good reception and especially benefits the fringe-area listener.

The over-all selectivity characteristic of a typical inexpensive table model receiver, shown in Fig. 3, indicates that the 3-db high-frequency audio cutoff is only 2.6 kc. Laboratory tests indicate that the receiver, when used for CSSB reception, may be detuned 2 or 3 kc, improving the high-frequency response. Figs. 16 and 17 confirm this improvement in high-frequency response.

Fig. 16—Frequency of typical CSSB system as a function of receiver detuning. Logarithmic frequency scale.

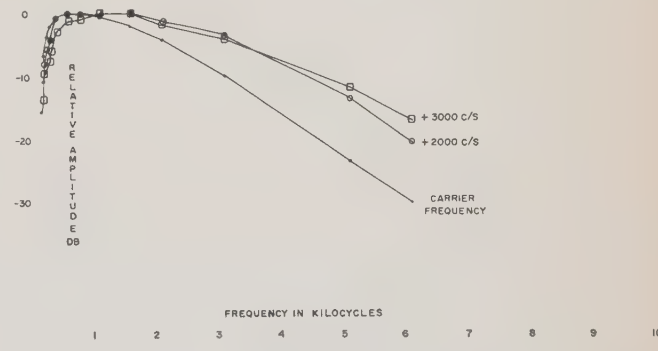


Fig. 17—Frequency response of typical CSSB system as a function of receiver detuning. Linear scale.

Both of these figures are plots of the over-all audio response of the receiver when tuned for AM (marked "carrier frequency" on the curve) or for CSSB with detuning of 2 and 3 kc. Fig. 17 is the same as Fig. 16 except that a linear plot which tends to emphasize the high-frequency characteristics of the response, is used instead of a logarithmic plot. The improved high-frequency response is especially noticeable in fringe-area listening where speech is unintelligible when AM is used.

It should also be noted that the amount of receiver detuning is a function of the signal-to-noise and signal-to-interference ratios. The smaller amount of detuning would occur when the receiver was operated under good signal-to-noise and interference situations. When the noise is high, or the interference is severe, the optimum amount of detuning increases. There are a number of reasons for this increase in allowable detuning.

The amount of detuning that the normal listener would tolerate is a function of the amount of distortion he hears. Under poor signal-to-noise situations, the noise masks distortion; therefore, the listener naturally tends to tune further from carrier in order to obtain increased volume and increased high-frequency response. This improved frequency response is extremely important under low signal-to-noise situations because the noise tends to mask the weak high-frequency sounds. Therefore, the average listener will, without being told, detune the receiver further in poor signal-to-noise situations than when the signal-to-noise is good.

IX. Co-Channel Interference

In considering co-channel interference, assume first the use of a receiver which has a perfectly flat pass band with an infinite cutoff slope IF/RF selectivity characteristic, and consider the case of two CSSB signals operating at the same carrier frequency but on opposite sidebands. Since a perfect receiver is considered, the cross-talk interference between these two co-channel stations would be equal to the undesired sideband radiation of the transmitter. Thus, if a 30-db desired-to-undesired sideband ratio were maintained at the transmitter, the isolation between the channels would be 30 db for equal signal strengths. If the geographical spacing provided additional isolation, then the isolation would, of course, be equal to 30 db plus the geographical isolation.

Consider the improvement that would be obtained when a CSSB signal operated simultaneously on the same channel as a conventional double-sideband AM signal. In this case, if the receivers had perfectly flat-top, sharp-sided selectivity characteristics, the double-sideband signal would be reduced by a factor of approximately 7 db when the CSSB wave was received. This loss is due in part to the fact that one of the sidebands of the interfering double-sideband signal is not accepted by the receiver and in part to the quadrature relationship between the carrier and remaining sideband. Therefore, the AM of the AM signal would be reduced to a little less than 50 per cent.

Listeners tuned to the double-sideband signal could, by using a perfect selectivity characteristic receiver, drop interference from the other station by 30 db. But, because his highly selective receiver would eliminate one of the desired sidebands, the desired modulation would be reduced by 7 db, and 11 per cent envelope distortion would be introduced (because of sideband wave). Thus, the AM wave listener would have a net gain in signal-to-interference ratio of 23 db, but at the same time considerable distortion would be introduced. Of course, he could still tune to the carrier and maintain the exact receiving conditions that existed when both stations were transmitting conventional double-sideband AM waves.

In practice, the listener automatically tunes his set to accommodate prevailing conditions. For instance, if his home is near the double-sideband AM station, he will normally tune to the carrier so as to receive the loudest signal. If, however, he lives near the interfering CSSB co-channel station, he will tune his set away from the sideband of the CSSB station so as to minimize interference. Thus, because the interfering station radiates only one sideband, the listener has the possibility of reducing interference by tuning away from the sideband used by the co-channel station. The CSSB signal listener can reduce interference effects by tuning away from the carrier and toward the desired sideband.

Some will question the ability of the average listener to optimize reception in this manner, but while every listener will not do this, most listeners, by tuning for the best sound, should approximate the desired result. This is especially true for listeners who live in areas where radio reception is poor, and are therefore accustomed to listening to marginal signals.

If both stations utilized the CSSB system and transmitted opposite sidebands, the greatest gain would be achieved. It should be noted that if the two CSSB co-channel stations transmit on the same sideband, there is neither improvement nor degradation in the interference situation. Of course, the other advantages of CSSB operation are still applicable.

The selectivity characteristics of common home receivers are, as mentioned above, far from perfect. (It should be noted that it is practical to make receivers with excellent selective characteristics which could provide almost the full interference improvement; however, in this study it is more important to consider sets now in use.) Actual laboratory tests which used conventional inexpensive receivers and which were made during the study of co-channel interference, verify the above conclusions. Figs. 18 and 19 show various conditions of co-channel interference under situations in which CSSB is used by the desired station or by both stations. A maximum gain of 10 to 15 db is indicated for the weak high AF components.

Another advantage of the CSSB signal is that it is slightly less sensitive to phase rotation of the carrier, especially at low percentages of modulation. This is because the other sideband is absent; therefore, since only two strong components are present at low percentages of modulation, the phase of the carrier does not influence the envelope waveshape for these low percentage modulation conditions. When the undesired signal has

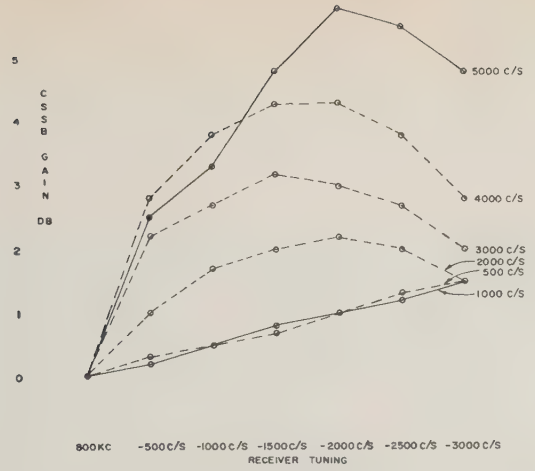


Fig. 18—Co-channel interference gain by use of CSSB as a function of receiver tuning. Transmitter creating interference operates AM, desired operates CSSB.

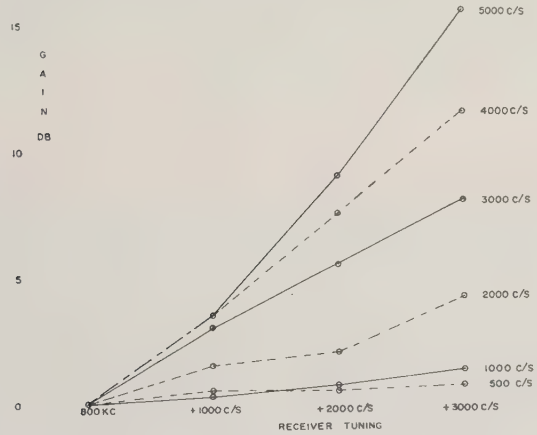


Fig. 19—Co-channel interference gain by use of CSSB as a function of receiver tuning. Transmitter creating interference and desired operates CSSB.

a carrier frequency approximately equal to that of the desired carrier frequency, the combined carriers will be phase modulated at a low-frequency beat-note rate. Because CSSB is less sensitive to phase discrepancies, it should be a little less sensitive to this form of carrier beating distortion.

X. Adjacent Channel Interference

In the following, it will be assumed that, except where noted, a 10-kc channel separation is used. The "shaped noise" technique, described above, was used throughout these measurements on adjacent channel interference. Tests with pure random noise modulation (up to 20 kc) indicate a much greater gain for CSSB; however, while strong high-frequency sounds are occasionally emitted, it is believed that the shaped noise test gives a far more realistic answer.

The four possible methods of arranging the sidebands between two interfering adjacent channel CSSB stations are shown in Fig. 20. Three of these arrangements ap-

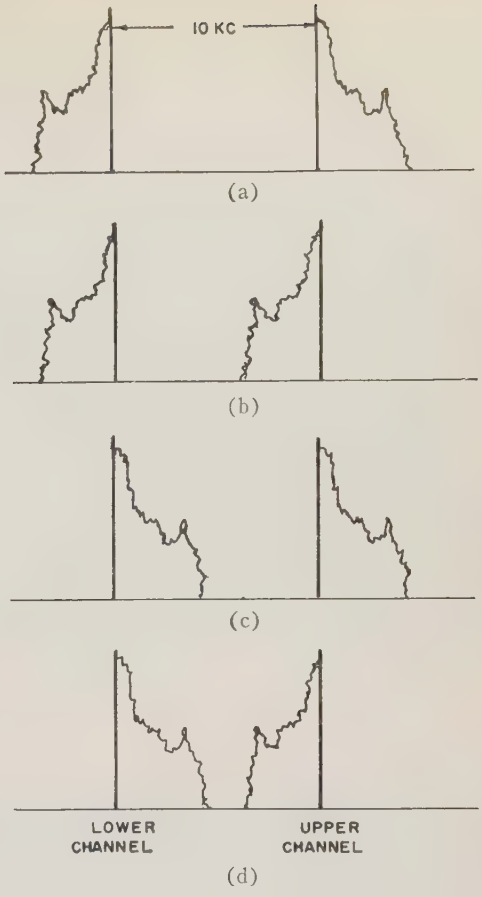


Fig. 20—Various conditions of adjacent channel interference.

pear to offer appreciable improvements in interference. The fourth would not greatly alter the interference situation by changing to CSSB operation.

In the first arrangement, Fig. 20(a), the higher frequency station utilizes the upper sideband and the lower frequency station takes the lower sideband. In this most favorable case, the nearest strong component would be at least 13 kc away from the center tuning point of the receiver. This 13-kc figure is arrived at by recognizing that the carriers are spaced by 10 kc and that the receiver, under strong interference conditions, would be tuned approximately 3 kc on the desired side. Fig. 21(b) shows actual measurements for this situation, and Fig. 21(a) shows the conventional AM situation which may be used for a comparison. It should be noted that detuning by 3000 cycles produces a gain for CSSB of 17 db. [See pp. 1520–1522 for Figs. 21(a)–(i).]

Fig. 20(b) shows the situation where both CSSB adjacent channel stations operate on the lower sideband. In this situation, the person listening to the lower frequency station would tune his receiver 3 kc below the carrier of that station. Therefore, the closest interference component would be approximately 5 kc from the center of the receiver's band pass. Fig. 21(c) indicates that under such conditions a gain of a little over 11 db would result.

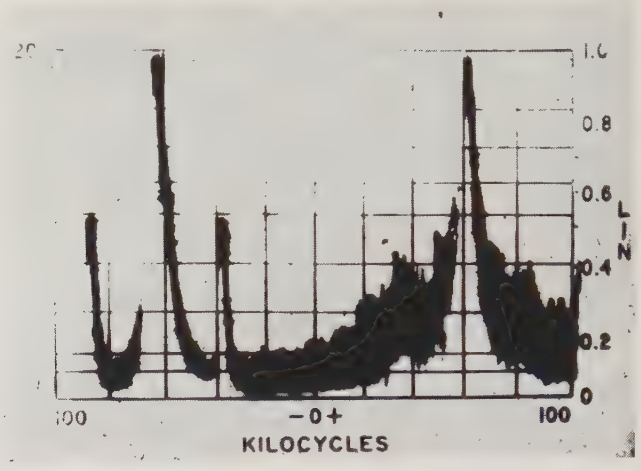
Adjacent Channel Interference
As a Function of Tuning
Using RCA Model 9-C-7EE Receiver
KRL Stereo Adapter
KRL CSSB Adapter
Two Experimental 5 Watt Test Transmitters

Desired Interference

AM/AM

Desired—AM 20 per cent at 2000 cps
Interference—Shaped noise AM at 100 per cent
Carrier amplitudes at equality with 10 kc separation

Signal to Interference Ratio +18.0 db

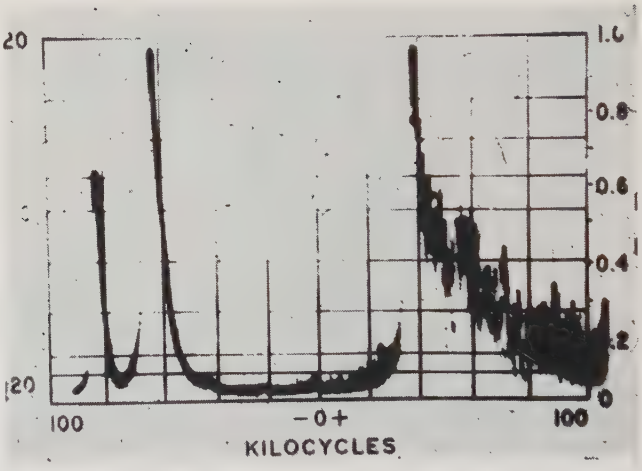


(a)

Lower Sideband/Lower Sideband

Desired—LSB 20 per cent modulation at 2000 cps
Interference—LSB shaped noise at 100 per cent modulation
Carrier amplitudes at equality with 10 kc separation

Receiver Tuning	Signal-to-Interference Ratio
Carrier	+19.5 db
−1000	+24.5 db
−2000	+27.5 db
−3000	+29.5 db



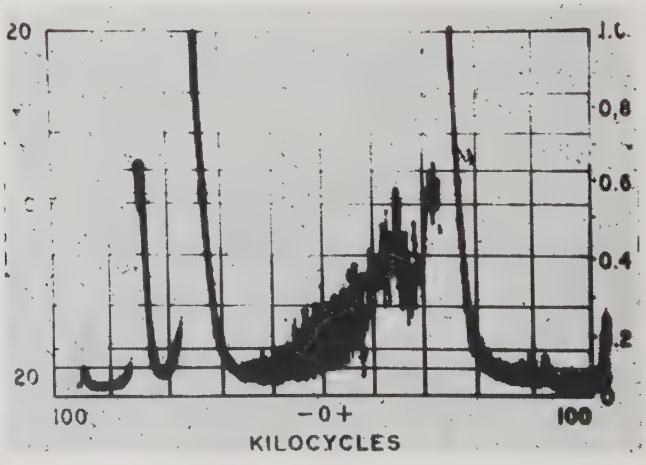
(b)

Lower Sideband/Upper Sideband

Desired—LSB 20 per cent modulation at 2000 cps
Interference—USB shaped noise at 100 per cent modulation
Carrier amplitudes at equality with 10 kc separation

Receiver Tuning	Signal-to-Interference Ratio
Carrier	+28.0 db
−1000	+31.5 db*
−2000	+33.5 db
−3000	+35.0 db*

* Measurement limited by ambient noise.



(c)

Fig. 21—Measurements for adjacent channel interference.

Adjacent Channel Interference

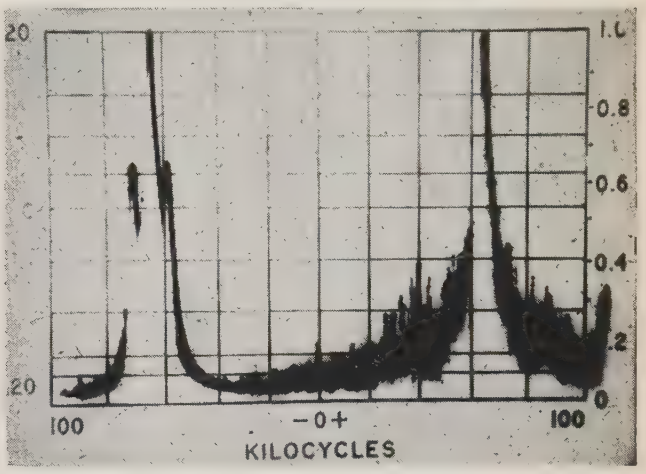
Using RCA Model 9-C-7EE Receiver
Measurements Signal Generator Model 65-B (as AM source)
KRL Stereo Adapter
KRL Compatible Adapter
Two Experimental 5 Watt Test Transmitters

Desired Interference

AM/AM

Desired—AM 30 per cent at 400 cps
Interference—Shaped noise AM at 100 per cent
Carrier amplitudes at equality with 10 kc separation

Signal to Interference Ratio +27.0 db

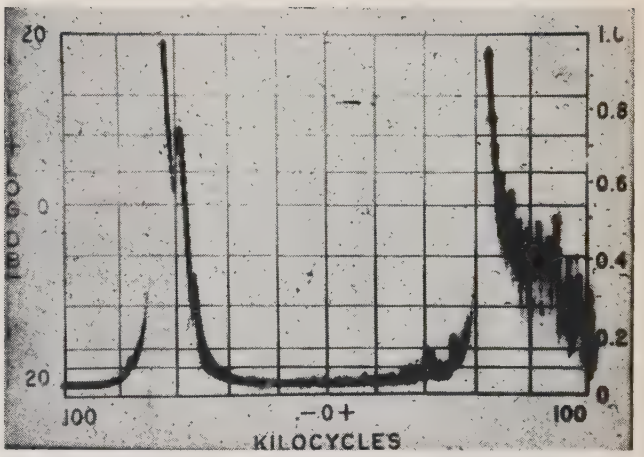


(d)

Upper Sideband/Upper Sideband

Desired—USB 30 per cent modulation at 400 cps
Interference—USB shaped noise at 100 per cent modulation
Carrier amplitudes at equality with 10 kc separation

Signal to Interference Ratio +37 db

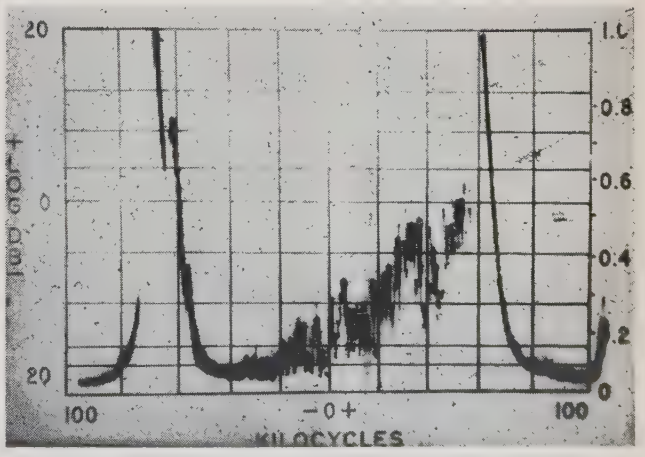


(e)

Upper Sideband/Lower Sideband

Desired—USB 30 per cent modulation at 400 cps
Interference—LSB shaped noise at 100 per cent modulation
Carrier amplitudes at equality with 10 kc separation

Signal to Interference Ratio +28 db



(f)

Fig. 21 (Continued)

Adjacent Channel Interference

Using RCA Model 9-C-7EE and RCA X1 Receivers
KRL Stereo Adapter
Experimental 5 Watt Test Transmitter (as CSSB source)
Measurements Signal Generator Model 65-B (as AM source)

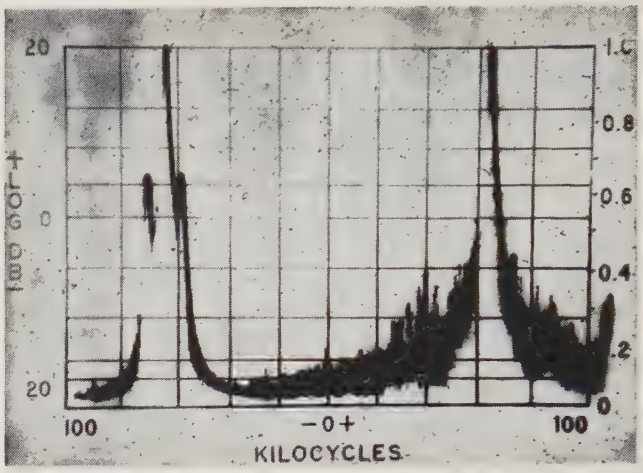
Desired Interference

AM/AM

Desired—AM 30 per cent at 400 cps
Interference—Shaped noise AM at 100 per cent
Carrier amplitudes at equality with 10 kc separation
Interfering carrier amplitude 30 times desired with 20 kc separation

Signal to Interference Ratio

10 kc	9-C-7EE	+27.0 db
	RCA X1	+31.0 db
20 kc	9-C-7EE	+ 4.5 db
	RCA X1	+14.0 db



(g)

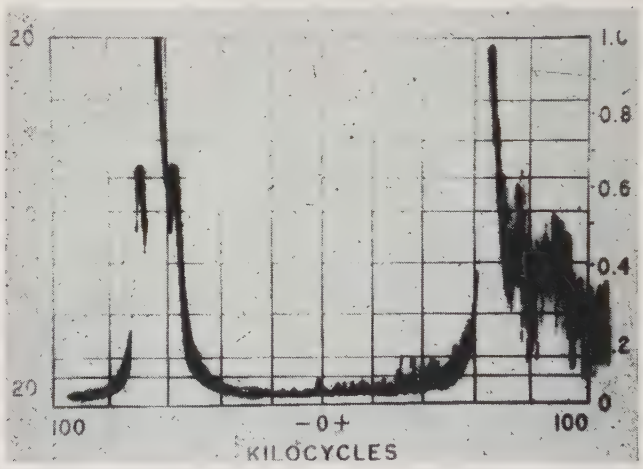
AM/Upper Sideband

Desired—AM 30 per cent at 400 cps
Interference—Shaped noise USB at 100 per cent modulation
Carrier amplitudes at equality with 10 kc separation

Signal to Interference Ratio

10 kc	9-C-7EE	+37 db
	RCA X1	+43 db*
20 kc	9-C-7EE	+33 db
	RCA X1	+44 db*

* limited by ambient noise



(h)

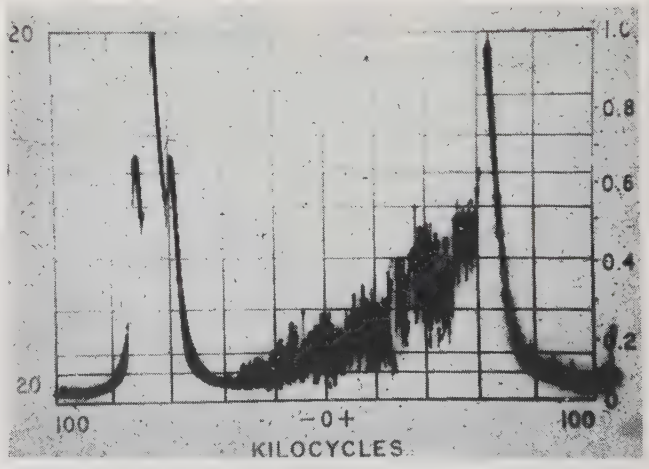
AM/Lower Sideband

Desired—AM 30 per cent at 400 cps
Interference—Shaped noise LSB at 100 per cent modulation
Carrier amplitudes at equality with 10 kc separation

Signal to Interference Ratio

10 kc	9-C-7EE	+28 db
	RCA X1	+35 db
20 kc	9-C-7EE	+33 db
	RCA X1	+44 db*

* limited by ambient noise



(i)

Fig. 21

In the next possible arrangement of sidebands, Fig. 20(c), the upper sideband is transmitted by both stations. In this instance, the person listening to the higher frequency station tunes his receiver approximately 3 kc above the carrier frequency, and the lower frequency station listener tunes his receiver to the carrier frequency. For the low-frequency listener, the nearest component would be 10 kc away from the center of the receiver's pass band. For the high-frequency listener, the nearest component would be some 5 kc away from the center tuning. The 5-kc figure is based upon the assumption that the listener will detune his receiver by 3 kc and that the crystal filter in the sideband transmitter adapter limits the strong signal components to 8 kc. The spectrogram shown in Fig. 21(e) provides measurements for this situation. The receiver is assumed to be tuned to the carrier of the low-frequency station and a gain of some 10 db over the AM situation has been measured. The case in which the listener tunes to the higher frequency station is not indicated in these figures but is identical to the spectrum situation shown in Fig. 21(c). In this case, a 1.5-db gain is provided when the receiver is tuned to the carrier, and over 11-db gain when the receiver is detuned 3 kc for best interference avoidance.

In the fourth situation, Fig. 20(d), the high-frequency station uses the lower sideband and the lower frequency station the upper sideband. This is the least desirable situation, and in this case the receivers should be tuned to the carriers for best reception in regions experiencing interference. The one advantage to the system is that the high-frequency components from the signal are restricted to 8 kc, whereas in present operation they are not restricted. There should be some improvements, therefore, in reception because of the foregoing factor. It is felt, however, that this advantage would be almost offset by the fact that the sideband level would be raised in both cases by 6 db. The spectrogram, Fig. 21(f), provides actual measurements for this situation, in which a small gain (1 db) was achieved for this CSSB situation over the conventional AM arrangement.

XI. Mixed CSSB and AM Transmission

It is obvious that, even if CSSB operation were accepted simultaneously in all countries, the transition to this form of operation would take many years; it is therefore very important to consider the situation of mixed CSSB-AM operation.

In the first case, assume that the sideband of the CSSB wave is away from the AM wave (if the CSSB were used by the lower frequency station, the lower sideband would be emitted; if it were used by the higher frequency station, the upper sideband would be emitted). The listener to the CSSB station, by detuning his receiver 3 kc, would gain approximately 7–10 db over AM operation. The AM station listener would achieve a gain of 10 to 12 db [see the spectrogram of Fig. 21(h)].

If the CSSB station picked the sideband closest to the AM wave, the CSSB signal listener would tune to the carrier, and there would be neither loss nor gain in interference. Also, the AM wave listener would have no real gain or loss in interference. The reason is that he would lose approximately 6 db because of the increased sideband levels from the CSSB signal, but would also gain, because the sideband components would be restricted and would not be closer than 2 kc to his carrier. Actual measurements [see Fig. 21(i)] indicate a 1 to 4 db gain for CSSB for 10 kc carrier separation.

XII. Detailed Consideration of Least Favorable Interference Situation for CSSB Operation

In proposing any new system, thorough consideration should be given to the least favorable operating situation, and for CSSB operation this would occur when the transmitted sideband was closest to the signal being disturbed. Even under this condition, there would be a small improvement over present AM operation for both stations' listeners.

The CSSB equipment specification limits the response to 8 kc. This specification was formulated because it is apparent from spectrum analysis of CSSB that some increase in interference for the worst condition can be expected in the absence of the 8-kc filter.

If one studies Fig. 16 or Fig. 17, it is apparent that an appreciable amount of high-frequency fall-off may occur in the CSSB transmission equipment while still providing improved over-all high-frequency audio characteristics.

Because of the 8-kc high-frequency response restriction, there is a gain of 1 to 4 db in the least desirable adjacent channel interference condition. [See Fig. 21(i).] As stated above, this is not a preferred condition of operation, and there is a much greater gain when the other sideband, the one farthest away from the station suffering interference, is used. However, there will necessarily be instances in which this condition will arise: thus the necessity for limiting the high-frequency response of the equipment.

With reference to measurements made on the interference characteristics of the CSSB system when the CSSB signal is spaced 20 kc from the desired signal, the improvement is quite substantial no matter which sideband is used by the CSSB broadcaster. For 20-kc adjacent channel interference, a gain of up to 28.5 to 30 db was measured over AM operation no matter which sideband was used. The reason for this gain is the use of the sharp sideband filter which insures that the bulk of the energy transmitted is restricted to a bandwidth of 8 kc. However, a note of caution is required. The interference characteristic for 20-kc carrier separation is extremely sensitive to the amount of high-frequency content in the modulating wave. Thus, since these measurements are highly dependent on the type of signal being transmitted, it is difficult to state unequivocally that in practice the full 20-kc separation gain will be achieved.

XIII. Television Receiver Radiation Interference Reduction

One of the most serious problems that presently faces the AM broadcaster is the interference produced by poorly shielded television receivers. This problem arises from the fact that most horizontal sweep frequency amplifiers have sufficient harmonic radiation to extend over the entire broadcast range and even up into the high-frequency range. This effect is similar to having some 68 high-powered broadcast stations covering the U. S. added to the broadcast band.

In 1958, questionnaires were sent by Kahn Research Laboratories, Inc., to some 1500 individuals in the New York City metropolitan and suburban area. The addresses were chosen by taking random columns from various telephone books. Ten replied that they did not have radio receivers, 205 stated that the television interference was not objectionable, and 211 said it was objectionable. Thus, over 50 per cent of those who actually answered the questionnaires indicated that they suffered from this type of interference. The importance of this problem has been confirmed by discussions with various broadcasters throughout the U. S. and Canada. One solution to this problem would be to provide better shielding in the television receiver, but this would increase the cost of the receiver. It would be unrealistic to expect that sets now in the hands of the public would be so altered.

The use of CSSB, however, does provide a solution to this problem for many broadcast channels. First of all, the very fact that CSSB increases the level of the high-frequency components of the desired signal should improve the signal-to-interference ratio. Also, by choosing the sideband away from the interference component, an appreciable gain can be achieved.

Fig. 22 is a chart which shows the frequency distribution of the television interference components. This chart was prepared by merely calculating the frequency of the harmonics of the 15.75-kc horizontal output frequency that fall in the broadcast band.

A study of this chart will show that in most cases an appreciable improvement can be achieved by selecting the proper sideband. For instance, if the station is operating on 1180 kc, it should select the lower sideband in order to reduce the television interference which falls at 1181.25 kc. The listener, by tuning his receiver for loudest sound, would reduce this form of interference.

In order to confirm these theoretical conclusions, various interference situations were checked with an average receiver tuned for optimum reception of CSSB when transmitting on the appropriate sideband for avoidance of this interference. (See Figs. 23 and 24.)

The various conditions of interference separation were investigated while varying the tuning of the receiver. It can be seen from Figs. 23 and 24 that the advantage is quite impressive (up to some 16 db) for large amounts of detuning and large separation of interference from the carrier.

Interference Frequency	*Interference Frequency*
551.25	1086.75
567	1102.5
582.75	1118.25
598.5	1134
614.25	1149.75
630	1165.5
645.75	1181.25
661.5 (42nd harmonic of 15.75 kc)	1197
677.25	1212.75
693	1228.5
708.75	1244.25
724.5	1260
740.2	1275.75
756	1291.5
771.75	1307.25
787.5	1323
803.25	1338.75
819	1354.5
834.75	1370.25
850.5	1386
866.25	1401.75
882	1417.5
897.75	1433.25
913.5	1449
929.25	1464.75
945	1480.5
960.75	1496.25
976.5	1512
992.25	1527.75
1008	1543.5
1023.75	1559.25
1039.5	1575
1055.25	1590.75
1071	1606.5 (102nd harmonic of 15.75 kc)

Fig. 22—Distribution of television interference in standard radio broadcast range.

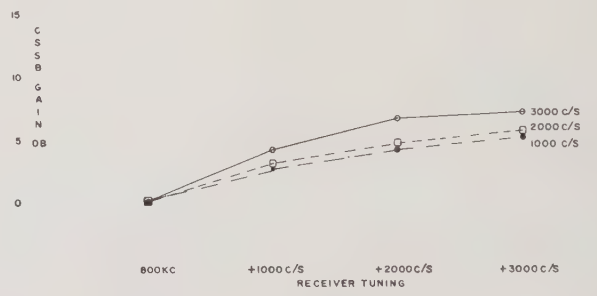


Fig. 23—CSSB gain in reducing TV interference as a function of receiver tuning. Desired modulation frequency 400 cycles. Curve markings indicate difference between carrier and interference frequencies.

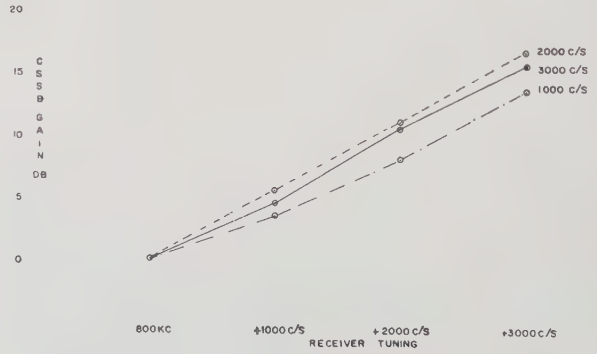


Fig. 24—CSSB gain in reducing TV interference as a function of receiver tuning. Desired modulation frequency 3000 cycles. Curve markings indicate difference between carrier and interference frequencies.

XIV. Reduction of Certain Types of Fading Distortion

Because of the lack of sensitivity to the phase of the carrier at low percentages of modulation, a reduction of fading distortion might be expected if CSSB were used. Laboratory tests, however, are unwieldy and inconclusive. Further proof would necessitate extensive field testing under actual broadcast operating conditions. It is therefore only possible to theorize that there may be some small advantage in the use of CSSB under fading distortion conditions.

XV. Measurements of Sensitivity of CSSB System to Receiver Tuning

One of the main questions that should be considered in evaluating the CSSB system is the question of sensitivity of the system to receiver tuning error.

Figs. 25–29 provide information regarding the tuning error sensitivity. Figs. 25–27 show total harmonic distortion of the entire system for two typical receivers when a 1000-cycle tone is used to modulate the CSSB transmitter, and its associated AM transmitter by 95 per cent. As a control for the test, a curve is provided showing the distortion produced when AM transmission is used. It should be stressed that in all cases over-all system distortion figures—that is, the distortion introduced by the CSSB adapter plus the transmitter and the associated receiver—are shown.

Fig. 26 shows a similar curve, but the equipment is modulated by 3 kc at 95 per cent modulation instead of by 1000 cycles. In all cases, the upper sideband is transmitted when the CSSB system is used. Fig. 27 shows

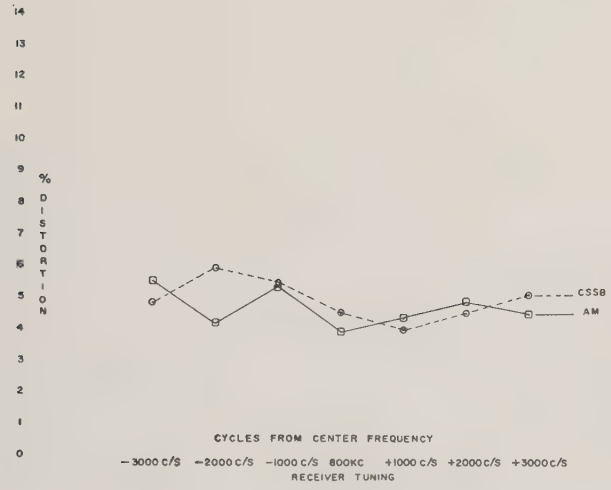


Fig. 25—Total harmonic distortion at output of receiver for AM and CSSB as a function of receiver tuning. Receiver used RCA Model X1. Modulation 1000 cycles (95 per cent).

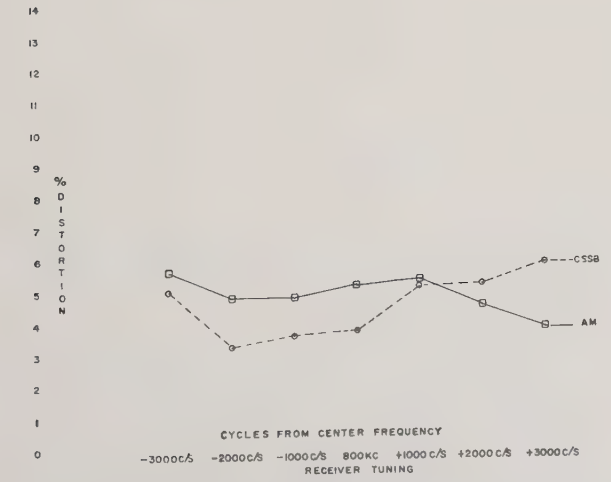


Fig. 26—Total harmonic distortion at output of receiver for AM and CSSB as a function of receiver tuning. Receiver used RCA Model X1. Modulation 3000 cycles (95 per cent).

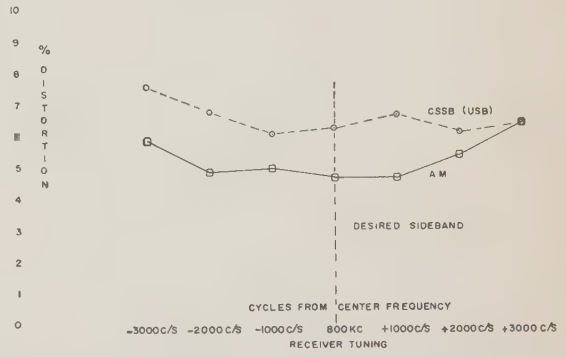


Fig. 27—Total harmonic distortion at output of receiver for AM and CSSB as a function of receiver tuning. Receiver used RCA Model 9C7EE. Modulation frequency used is 1000 cycles (95 per cent).

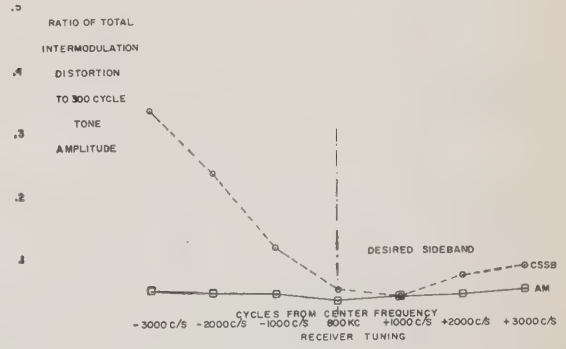

RATIO OF TOTAL INTERMODULATION DISTORTION TO 300 CYCLE TONE AMPLITUDE
DESIRED SIDEBAND
CSSB
AM
CYCLES FROM CENTER FREQUENCY
-3000 C/S -2000 C/S -1000 C/S 800KC +1000 C/S +2000 C/S +3000 C/S
RECEIVER TUNING

Fig. 28—Over-all system intermodulation distortion as function of receiver detuning. RCA Model X1. Modulation frequency 300 and 3000 cycles peak percentage modulation 85 per cent.

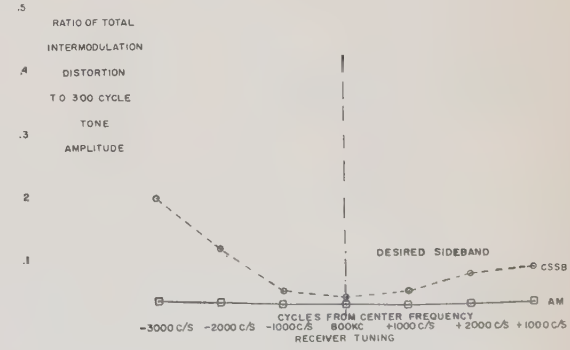

RATIO OF TOTAL INTERMODULATION DISTORTION TO 300 CYCLE TONE AMPLITUDE
DESIRED SIDEBAND
CSSB
AM
CYCLES FROM CENTER FREQUENCY
-3000 C/S -2000 C/S -1000C/S 800KC +1000C/S +2000 C/S +1000C/S
RECEIVER TUNING

Fig. 29—Over-all system intermodulation distortion as function of receiver detuning. RCA Model 9C7EE. Modulation frequency 300 and 3000 cycles peak percentage modulation 85 per cent.

the harmonic distortion when a second receiver was tested.

Figs. 28 and 29 provide information regarding intermodulation distortion. A test signal was used which included a 300-cycle tone and a 3000-cycle tone, the 3000-cycle tone having one-fourth the amplitude of the 300-cycle tone. The peak modulation used for this test was 85 per cent.

These tests indicate that while the harmonic distortion of CSSB is quite low for a fair amount of detuning, the intermodulation distortion rapidly rises when one tunes to the attenuated sideband. It should be noted, however, that the listener would know that he was tuned to the incorrect sideband because the output level would be very low.

While these measurements indicate that the system may be less tolerant to incorrect tuning, an improperly tuned receiver is obvious to the listener because of the decreased volume. The incorrect tuning is easily and naturally corrected, and there is a compensating factor. By far the greatest trouble experienced by the public when tuning a receiver is caused by interference from adjacent or co-channel transmissions or from television receiver radiation interference. By reducing these difficulties, the system decreases the problem of tuning. Thus, it is felt that the sensitivity in tuning is more than compensated by the reduced difficulty in tuning for avoidance of interference.

XVI. Results of Limited On-the-Air Tests

A number of stations participated in tests of CSSB, including KDKA, Pittsburgh;[21] WGBB, Freeport, Long Island, N. Y.; WMGM, New York; WABC, New York; and WSM, Nashville. In addition, there are a number of U. S. Government and foreign stations in Europe, South America, and Asia, as well as in Canada, equipped for CSSB operation.

As to the increased level and improved fidelity, only one of the stations that conducted CSSB tests requested listener comments. This station was WGBB in Freeport, a 250-watt station (the only low-powered station which participated in these tests). According to their report, their solicitation resulted in the following listener comments:

Improved reception with CSSB	118,
No difference in signal	26,
Poorer reception with CSSB	7.

Some of the 118 favorable comments included: "better than ever," "very loud and clear," "no more interference from that Brooklyn station, now very clear even on my old radio," "previous to installation of new equipment we hadn't been getting WGBB clearly," "sound is much better, in fact I don't believe that it could be any better. I used to have trouble tuning in," "WGBB tone and fidelity has definitely been improved. Particularly noticeable in car radio on North Shore of Long Island where reception had previously been poor, or none at all."

At most of the CSSB experimental broadcast stations, the equipment achieved an average of 30 db of sideband rejection. These figures, for high-powered operation (many test stations operated at a 200 kw peak envelope power rating), compare quite favorably with standard high-power SSB transmitters. A number of different types of transmitters were used, including Doherty type, high-level, and low-level systems. At several of these stations, the equipment was used without readjustment for some three months of operation, and at the end of these test periods the sideband suppression still exceeded the 30-db figure. Thus, it is felt that satisfactory CSSB operation can be achieved with little required maintenance.[21] It should be noted that in all of these cases, the 1.4 phase stretch circuit was used, and it is felt that improved performance could be achieved by use of the newer and simpler 0.5 PM plus 1 PM circuit as this new circuit avoids the use of a critical frequency divider circuit and can also withstand reasonable amounts of overmodulation.

Tests to determine the suitability of CSSB transmission for aeronautical service were made by Aeronautical Radio, Inc., which is a communications company specializing in aeronautical mobile service. It is owned principally by the airlines. These tests, which ran approximately three months, transmitted CSSB from the Long Island, N. Y., ARINC station to conventional airborne AM receiving equipment flying in the Caribbean area, as well as to a military aircraft equipped with a suppressed carrier sideband receiver. The tests were quite successful and Aeronautical Radio has recommended CSSB as one of the four acceptable SSB techniques for ground-to-air and air-to-ground service.[22]

XVII. Summary of Advantages of the CSSB System

The advantages of the CSSB system over conventional amplitude modulation for broadcast service may be summarized as follows:

1) A signal-to-noise power gain for a *given audio fidelity*. This may be seen by recognizing the fact that the bandwidth of a receiver for CSSB need only be $\frac{1}{2}$ of that for AM. Therefore, white noise should be reduced by 3 db and impulse noise by

[21] Westinghouse reported to the Federal Communications Commission that not a single complaint of bad reception of any type was received from KDKA's transmissions on CSSB for a period of over five months.

The chief gain of using CSSB to KDKA is the reduction of adjacent channel interference between KDKA on 1020 kcs and WBZ in Boston on 1030 kcs. The removal of KDKA's upper side frequencies has relieved interference to WBZ's reception and the ability of receivers tuned to KDKA to tune slightly off resonance away from WBZ and toward the radiated lower sideband of KDKA has also given KDKA some additional immunity from WBZ.

In our experience with the Kahn adapter at KDKA it has been reliable, stable, and simple to operate.

We believe further experience will gain additional information which will probably show the Kahn adapter is most useful as an allocation tool.

[22] "Airborne HF SSB/AM System," ARINC Characteristic No. 533; February 15, 1960.

some 6 db. In practice, it appears that due to economics and the requirements for reduction of adjacent channel interference, the common broadcast receiver is actually more suitable for CSSB than for double-sideband AM reception.

2) Improved audio fidelity. Because, as mentioned above, the receivers are too narrow for AM, halving the bandwidth of the signal allows the higher frequency components to be passed, resulting in improved fidelity.
3) Reduction in co- and adjacent-channel interference: a) to the CSSB listener, and b) to the station which previously experienced interference from the station now using CSSB.
4) Provision of a means for reducing interference caused by improperly-shielded TV receivers and other devices, because, under many practical conditions, the station may choose the appropriate sideband to reduce the interference.
5) Possible reduction of selective fading distortion. There is very little evidence regarding this last advantage, and it is the author's feeling that this may turn out to be of relatively little importance in practice.

The advantages of CSSB for communications service are the same as above stated for broadcast service, plus the following:

1) In comparison with conventional SSB and in common with conventional AM, CSSB does not suffer from sensitivity to Doppler shift, because the carrier is transmitted.
2) CSSB can be used with existing receivers and generally with a much lower priced type of receiver than required for conventional SSB reception.
3) CSSB can enjoy full advantage of speech clipping[23,24] which provides some 6 to 9 db of gain and allows the equipment to be operated with less supervision. This is not true for conventional SSB.

On the other hand, in comparison with a fully implemented SSB system, CSSB is more sensitive to selective fading. Its main applications besides AM broadcasting appear to be in mobile telephone and other voice communications circuits where cost and equipment complexity are major factors.

[23] L. R. Kahn, "The use of speech clipping in single-sideband communications systems," PROC. IRE, vol. 45, pp. 1148–1149; August, 1957.
[24] W. Lyons, "SSB and ISB systems for long-distance radio telegraph," *Trans. AIEE*, vol. 46 (*Commun. and Electronics*), pp. 921–924; January, 1960.

Conclusions

Since, in the CSSB system, only the low-frequency voice and music components have sufficient amplitude to produce appreciable second-order sideband components, the bandwidth of the CSSB wave is essentially that of a conventional SSB wave.

Because this new type of wave occupies approximately one-half the bandwidth which an AM wave occupies, CSSB effectively reduces co- and adjacent-channel interference. Equipment was described which makes it possible to adapt conventional AM transmitters to this mode of operation. Measurements indicate that most of the theoretical advantages of this system are achieved in actual practice.

Acknowledgment

The development of the CSSB system has been an extremely long and involved project and the author wishes to acknowledge the assistance of many distinguished engineers.

First of all, the author would like to acknowledge the participation of his colleagues at Kahn Research Laboratories, including Lloyd A. Ottenberg, Robert C. Rogers, Heino L. Pull, Dr. Leonard O. Goldstone and Kenneth B. Boothe. In addition, he wishes to thank Ralph Harmon, George Hagerty, A. C. Goodnow and T. C. Kenney of the Westinghouse Broadcasting System; Gatien Dandois of the Canadian Broadcasting Company; William T. Carnes of Aeronautical Radio; Charles Affelder and Julius Ross of the United States Information Agency; Leucien E. Rawls, formerly of WSM; Arthur Gulbis of Radio Rumbos, Caracas, Venezuela; and Aurele Borsuirt of CJAD, Montreal, Quebec, Canada.

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Magnetic Circuits Employing Ceramic Magnets*

J. C. HELMER†

Summary—Measurements of the field produced by a ceramic, brick-shaped magnet show that the assumption of fixed, uniform magnetization is a useful approximation to the physical state of the magnet. One consequence is that the magnetization is easily measured without disturbing the physical state of the magnet. Another consequence is that the incremental permeability of a ceramic magnet is approximately that of free space. This leads to a significant simplification in the equations describing the fields produced in magnetic circuits employing ceramic magnets. Several types of magnetic circuits are analyzed from this new point of view. A method of design optimization is developed which, in one particular case, violates the ancient criterion that the magnet operate at the point of maximum *BH* product. A generalized derivation of the *BH* energy theorem is given. The properties of steel magnets are contrasted with those of ceramic magnets. In an effort to make use of the leakage field of steel horseshoe magnets, a low-leakage ceramic-steel magnet is developed in which both magnetic materials operate at their respective points of maximum *BH* product. A method is described and analyzed by which the uniformity of the field produced by ceramic pole pieces is improved by covering the pole faces with thin sheets of iron.

I. Introduction

IT IS sometimes stated that the design of permanent magnet circuits is more of an art than a science. This statement applies particularly to circuits containing steel magnets. The situation is quite different when we consider the properties of ceramic magnets. Here we have a material with fixed magnetization and a permeability equal approximately to that of free space. The resulting simplification in the field equations has not been pointed out in the literature. A consequence of this simplification is that it is possible for a novice to build a circuit out of blocks of iron and precharged blocks of ceramic magnet and end up with a precalculated field. Therefore, it is the purpose of this paper to show that the design of ceramic, permanent magnet circuits is more of a science than an art.

An important consequence of the above remarks is that it is not necessary for the manufacturer of a device employing a ceramic permanent magnet to rely upon the magnet maker for the design of the magnetic circuit. Thus, the device designer may employ a unified approach which optimizes all properties of the device. This work has been motivated, in particular, by problems arising in the design of VacIon®, high vacuum pumps which depend for their operation on strong, uniform, magnetic fields.

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II. The Theory of Permanent Magnet Circuits

A. The Field Potentials

Magnets are endowed with a magnetization M through which the fields they produce may be calculated. Depending on whether the source of M is in atomic currents or magnetic poles, we may have (in mks units) either

$$\boldsymbol{B} = \nabla \times \boldsymbol{A}, \qquad \boldsymbol{A} = \frac{\mu_0}{4\pi}\int \frac{\nabla \times \boldsymbol{M}}{r}\, dV \tag{1}$$

or

$$\boldsymbol{H} = -\nabla\Phi, \qquad \Phi = \frac{-1}{4\pi}\int \frac{\nabla\cdot\boldsymbol{M}}{r}\, dV, \tag{2}$$

where $\boldsymbol{A}$ is the vector potential resulting from a volume current density $\boldsymbol{i} = \nabla\times\boldsymbol{M}$, and Φ is the scalar potential resulting from a volume pole density $\rho = -\nabla\cdot\boldsymbol{M}$.[1] Since free magnetic poles do not exist (?), some writers prefer on the grounds of physical reality to work with the vector potential $\boldsymbol{A}$. However, for a given volume distribution of $\boldsymbol{M}$, either potential produces the same results, and the choice of which potential to use is dictated by the nature of the problem. Permanent magnet circuits frequently employ bodies of material (iron) which have nearly infinite permeability relative to the other parts of the circuit. In such bodies the approximation $\boldsymbol{H}=0$ is good and therefore the surface of such a body becomes an equiscalar potential. The fictitious magnetic poles move about on this surface in much the same way as electric charge on a conductor and the resulting field may be found by solution of Laplace's equation $\nabla^2\Phi = 0$ in the external space. For this reason the scalar potential Φ is most useful in determining the fields of permanent magnets.

For illustration, consider the example of a bar magnet with fixed, uniform magnetization, terminated on each end with an irregular blob of iron. Since the magnetization of the bar is fixed, the permeability of the bar is the same as that of free space. The magnetic field may be found by solving Laplace's equation subject to the condition that each blob of iron defines an equipotential surface. Afterwards the potential difference is adjusted so that the number of poles on the surface of each blob (including the surface in contact with the magnet) is the same as that which previously existed

[1] W. K. H. Panofsky and M. Phillips, "Classical Electricity and Magnetism," Addison-Wesley Publishing Co., Reading, Mass., ch. 8; 1955.

on the ends of the bar magnet. Or, in cgs units,

$$4\pi \int_{\text{magnet end}} \boldsymbol{M}\cdot d\boldsymbol{A} = \int_{\text{iron surface}} \boldsymbol{H}\cdot d\boldsymbol{S}. \qquad (3)$$

The concept of fixed magnetization applies particularly well to ceramic magnets. This concept may be applied to steel magnets if we also assign to them a relative permeability greater than unity, as determined approximately by the slope of the demagnetization curve for values of $\boldsymbol{H}$ between zero and the point of maximum energy product. In this case the behavior of the magnet is analagous to that of a dielectric with self-polarization, and the additional boundary defined by the surface of the permanent magnet greatly increases the difficulty of obtaining analytic solutions for the magnetic field.

B. The $B\cdot H$ Integral Theorem

Consider an arbitrarily selected volume V in a space throughout which $\nabla\cdot\boldsymbol{B}=0$ and $\nabla\times\boldsymbol{H}=0$. We will compute the $B\cdot H$ integral within this volume.

$$W_i = \int_V \boldsymbol{B}\cdot\boldsymbol{H}dV = -\int_V \boldsymbol{B}\cdot\nabla\Phi dV \qquad (4)$$

where Φ is the scalar potential for H.

Note, however, that

$$\nabla\cdot\boldsymbol{B}\Phi = \Phi\nabla\cdot\boldsymbol{B} + \boldsymbol{B}\cdot\nabla\Phi = \boldsymbol{B}\cdot\nabla\Phi. \qquad (5)$$

Then

$$W_i = -\int_V \nabla\cdot\boldsymbol{B}\Phi dv = -\int_S \Phi\boldsymbol{B}\cdot d\boldsymbol{S} \qquad (6)$$

where S is the surface of the volume in question and $d\boldsymbol{S}$ is a vector directed outward normally from this surface. We will compute now the $\boldsymbol{B}\cdot\boldsymbol{H}$ integral over all of the space outside of V. By the same argument we arrive at an expression of identical form

$$W_0 = -\int_S \Phi\boldsymbol{B}\cdot d\boldsymbol{S}_-, \qquad d\boldsymbol{S}_- = -d\boldsymbol{S}, \qquad (7)$$

where $d\boldsymbol{S}_-$ is a vector directed normally towards the interior of V. The integral over the boundary at infinity vanishes because at a sufficiently great distance r, the strongest type of field we can have is a dipole field whose strength goes as $1/r^3$. Then

$$\int \Phi\,|B|\,|dS| \sim \frac{1}{r} \to 0 \quad \text{as} \quad r\to\infty. \qquad (8)$$

Since Φ is single valued and $\boldsymbol{B}$ is continuous, we conclude that

$$-W_i = W_0. \qquad (9)$$

Therefore, the $\boldsymbol{B}\cdot\boldsymbol{H}$ integral within V is equal to minus the $\boldsymbol{B}\cdot\boldsymbol{H}$ integral outside of V. This can be only if $\boldsymbol{B}\cdot\boldsymbol{H}\equiv 0$ or if there is some region in which $\boldsymbol{B}$ and $\boldsymbol{H}$ components are oppositely directed. The interior of a magnet is such a region, as shown in Fig. 1.

The above integral theorem becomes useful when the surface S defines the boundary of a permanent magnet. Then the volume of the magnet is given by

$$V = \frac{W_0}{-(\boldsymbol{B}\cdot\boldsymbol{H})\ \text{average in magnet}}. \qquad (10)$$

For a given external-field energy the volume of magnetic material is minimized by maximizing the $-\boldsymbol{B}\cdot\boldsymbol{H}$ product in the magnet.

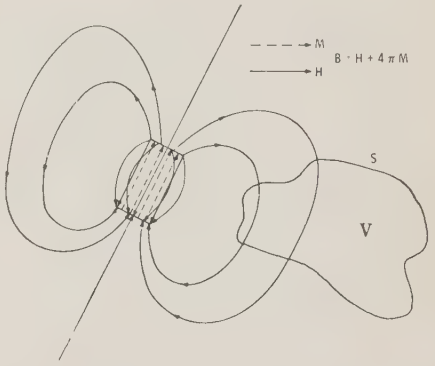


Fig. 1—Field of a permanent magnet.

Consider now a magnet with fixed, uniform magnetization in a uniform field H. We can write (in cgs units) $\boldsymbol{B}=\boldsymbol{B}_r+\boldsymbol{H}$, where $B_r=4\pi M$ is the magnetic remanence given by the value of B when $H=0$. Then

$$-\boldsymbol{B}\cdot\boldsymbol{H} = -(\boldsymbol{B}_r+\boldsymbol{H})\cdot\boldsymbol{H} = -B_rH\alpha - H^2, \qquad (11)$$

where α is the cosine of the angle between $\boldsymbol{B}_r$ and $\boldsymbol{H}$. Thus, the maximum value of $-(\boldsymbol{B}\cdot\boldsymbol{H})$ is obtained when

$$H = \frac{-B_r\alpha}{2}, \qquad -(\boldsymbol{B}\cdot\boldsymbol{H})_{\max} = \frac{B_r^2\alpha^2}{4}, \qquad (12)$$

and this, in turn, is maximized when $\alpha=-1$ so that $\boldsymbol{H}$ is antiparallel to $\boldsymbol{B}_r$. Thus, for a given external-field energy the volume of a ceramic magnet is minimized by designing the circuit so that within the magnet, H is antiparallel to M and has the value $H=B_r/2$. It may not always be possible to design the circuit so that the magnet operates in this way. If, as we shall see in Section IV, there are constraints which fix the cross section of the magnet, then the optimum point of operation is not at the point of maximum BH product.

In steel magnets the point of maximum BH product is set by the value of H at which severe demagnetization sets in. This may be seen in the demagnetization curves in Fig. 2. The demagnetization curves of magnets of equal maximum BH product are tangent to the same $BH=$const. curve. Thus, according to the BH energy theorem, there is not nearly as much difference in re-

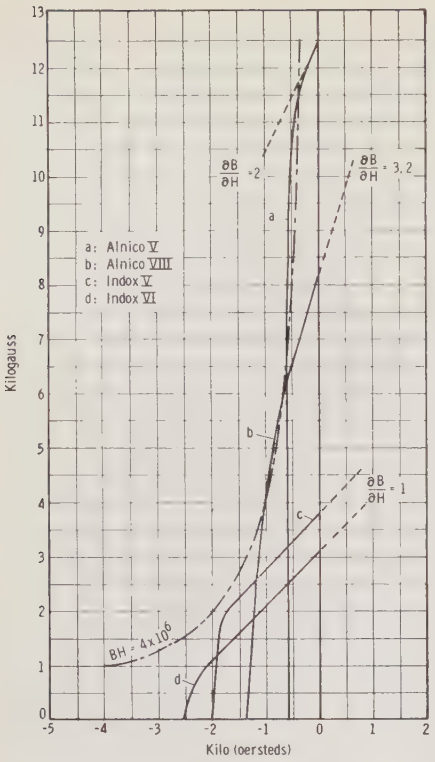


Fig. 2—Demagnetization curves.

quired volume between the ceramic and steel magnets as one might suppose from the difference in magnetic remanence.

C. Efficiency

Through (10) we may define the efficiency of a magnetic circuit as follows:

$$\text{Efficiency} = \frac{\text{measured gap energy} \times 2}{\text{magnet volume} \times (-B \cdot H)_{\max}}. \quad (13)$$

If in the gap energy we include only that part of the field which is useful, then (13) is a measure of the overall quality of the circuit. If in the gap energy we include the energy of leakage fields as well, then (13) is a measure of the efficiency of utilization of the permanent magnet material. It is difficult to estimate the leakage energy, although in certain simple geometries it is possible to use analytic methods.[2]

III. Ceramic Magnets

A. Demagnetization Curves

In Fig. 2 are shown the demagnetization curves for several types of magnetic materials supplied by the Indiana Steel Products Co. Two barium-ferrite ceramic magnets are shown, denoted by Indox V and VI. Over a wide range in H the demagnetization curves of the ceramic magnets are characterized by straight lines with slope $dB/dH \simeq 1$. If we plot $4\pi M = B - H$ against H, we get horizontal lines which show that the magnetization M is fixed and independent of H. Indox becomes demagnetized eventually, but this occurs beyond the point $H = -(B_r/2)$, so that over most of its working range it may be considered an ideal magnet. The incremental permeability of these magnets is given actually as approximately 1.05. This means that a demagnetizing field of strength $H = B_r/2$ will reduce the magnetization by about 2.5 per cent. The assumption of fixed magnetization, therefore, introduces a minimum error in the field calculations of about this size.

Ceramic magnets are frequently available in brick form, magnetized in a direction normal to one of the faces of the brick, as in Fig. 3. For uniform magnetization $\nabla \cdot \boldsymbol{M}$ is zero in the interior of the magnet, and is nonzero only at the faces normal to M. On these faces the divergence of $\boldsymbol{M}$ gives rise to a magnetic-pole density equal to M poles per unit area. The field arising from one surface layer alone is strong enough so that for Indox V the material just below the surface is on the verge of becoming demagnetized. This is seen by noting that the field adjacent to the surface layer, due to the surface layer alone is

$$H = -2\pi M = -\frac{B_r}{2}, \quad (14)$$

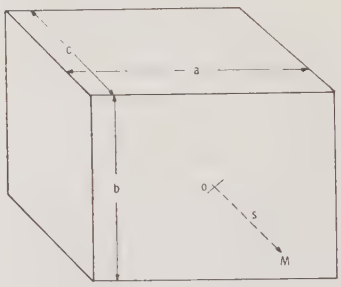


Fig. 3—Geometry of a brick-shaped magnet.

where B_r is the magnetic remanence. The presence of the opposite surface layer increases the internal demagnetizing field so that for samples such that $a, b \gg c$ in Fig. 3 the external field is cancelled and the internal field is $H = -B_r$. Such a condition will produce demagnetization in Indox V and VI magnets. This places restrictions on the dimensions of brick-shaped magnets if they are to be fully charged before being placed in the circuit. For instance, an Indox V magnet with dimensions $a = b = 3''$, $c = 2''$ became about 20 per cent demagnetized when separated into two pieces of dimensions $a = b = 3''$, $c = 1''$. To prevent demagnetization of Indox V, a practical rule is that c, the distance between pole faces, be at least two-thirds of the shortest face dimension. If this condition is not met, then the magnets may be charged in their circuit at a saturation flux density of 10,000 gauss.

[2] One reviewer suggests that the leakage energy may be obtained by subtracting the measured gap energy from the magnet energy.

B. Measurement of the Remanence of Ceramic Magnets

A method of measuring the remanence of a ceramic magnet is to measure the field strength along an axis s that passes through the center of the pole faces. For simple rectangular or cylindrical magnet geometries the field may be calculated and the remanence determined by matching the theoretical curve to the experimental points. This has been done for brick-shaped Indox V magnets with the following results.

The field due to one pole face of a brick-shaped magnet, as shown in Fig. 3 is given by

$$H = \frac{B_r}{4\pi}\int_{-b/2}^{b/2}\int_{-a/2}^{a/2}\frac{s\,dy\,dx}{(x^2+y^2+s^2)^{3/2}} = \frac{B_r}{\pi}\tan^{-1}\frac{ab}{2s(4s^2+a^2+b^2)^{1/2}}, \tag{15}$$

where a, b are the dimensions of the pole face, and s is the normal distance from the center of the pole face. If c is the face separation, then the total field is given by

$$H = \frac{B_r}{\pi}\left\{\tan^{-1}\frac{ab}{2s\sqrt{4s^2+a^2+b^2}} - \tan^{-1}\frac{ab}{2(s+c)\sqrt{4(s+c)^2+a^2+b^2}}\right\}. \tag{16}$$

The field at a distance s from one pole face on the axis of a *cylindrical* magnet, due to that pole face alone, is given by

$$H = \frac{B_r}{4\pi}\int_0^a \frac{2\pi s r\,dr}{(s^2+r^2)^{3/2}} = \frac{B_r}{2}\left(1-\frac{s}{(s^2+a^2)^{1/2}}\right), \tag{17}$$

where a is the radius of the cylinder.

In Figs. 4–6 are shown the results of measurement and calculation for three Indox V magnets of different dimensions. At distances greater than one inch from the pole face there is very precise agreement between theory and experiment. Thus, the remanence is quite accurately determined. Very close to the center of the pole face the field falls off somewhat below the calculated value. This indicates that the magnetization at the center of the block is a little below average.

One magnet was found in which the field strength at the face center was lower than the field strength a small distance from the face center. Measurements of total field strength along a perpendicular line through the axis at a constant distance from the face, as shown in Fig. 7, indicate that this effect is confined to a small region in the neighborhood of the face center. For this reason it is not a large source of error. In fact the field strengths obtained when these magnets are placed in circuits are close enough to the calculated values based on the measured remanence, that one suspects that the magnetization at the center has partly recovered itself

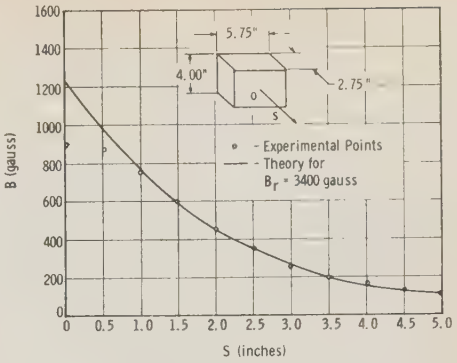


Fig. 4—Measured and calculated field of a ceramic magnet

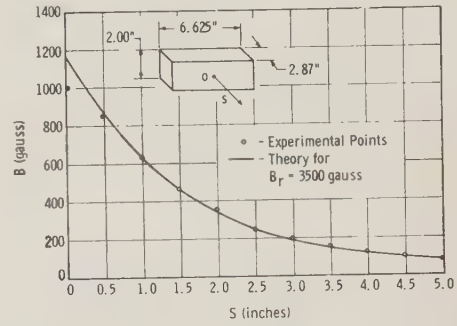


Fig. 5—Measured and calculated field of a ceramic magnet.

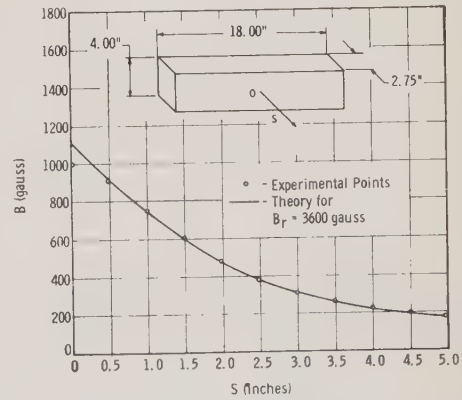


Fig. 6—Measured and calculated field of a ceramic magnet.

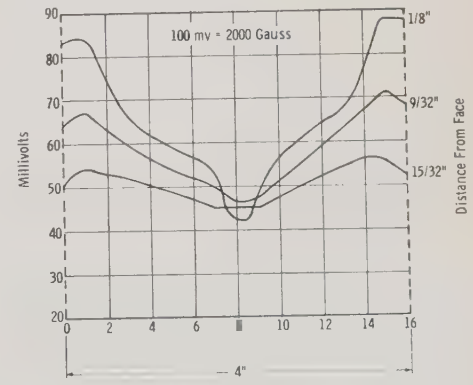


Fig. 7—Field profile on the face of a ceramic magnet.

under the influence of the new field distribution. The reason that an isolated magnet will become preferentially demagnetized at a face center is that the demagnetizing field of the opposite face is strongest at this point.

A feature of the theoretical field is that it falls off most rapidly close to the pole surface. This violates the intuitive conclusion, based on the behavior of fields adjacent to equipotential surfaces, that the slope of the field function in the normal direction at the surface should be zero. This conclusion is invalid because uniform charge surfaces are not equipotential surfaces. The behavior of fields close to uniform charge surfaces is clarified by the following argument. Consider the field component normal to the plane of a uniformly charged, planar surface of area A. The angle θ between the radius vector r, from an element of area dA on the surface to the field point, and the normal to the plane is the same as the angle between the radius vector r and the normal to dA, since dA is part of the plane. Thus, the normal field component is given by

$$H_n = \rho \int \frac{dA \cos\theta}{r^2} = \rho\Omega, \tag{18}$$

where $\rho = B_r/4\pi$ is the uniform charge density on the planar surface A and Ω is the solid angle subtended by the surface A at the field point. It is clear now that (15) and (17) are derived from solid-angle calculations. It is clear also that the solid angle Ω varies most rapidly with normal displacement adjacent to the charge plane.

IV. Magnetic Circuits Containing Ceramic Magnets

A. *Field of a Magnet in a Long Cylindrical Armature*

The field of two uniformly-magnetized ceramic pole pieces in a very long cylindrical armature, shown in Fig. 8(a), may be calculated exactly by image theory in two dimensions. Each line charge dq at radius r gives rise to an image charge $-dq$ in the same direction at radius $r' = R^2/r$, where $dq = B_r dx/2\pi$ and R is the inner radius of the armature. The field at the center is equal to twice the field produced by one pole face and its image alone. Thus, the field at the center is

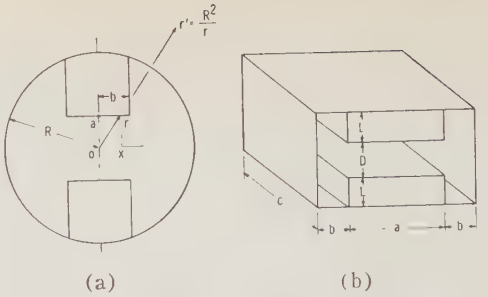
$$H = 2\int_{-b}^{b} \frac{B_r dx}{2\pi} \cdot \frac{a}{\sqrt{a^2+x^2}} \cdot \left(\frac{1}{\sqrt{a^2+x^2}} - \frac{\sqrt{a^2+x^2}}{R^2} \right) \tag{19}$$

$$\therefore\ H = \frac{2B_r}{\pi} \left\{ \tan^{-1}\left(\frac{b}{a}\right) - \frac{ab}{R^2} \right\}. \tag{20}$$

It is evident that the term $\tan^{-1} b/a$ arises from the pole faces alone, while the term $-ab/R^2$ arises from the image charges. In many practical designs the image term may produce only a small correction so that a useful estimate of the field strength, representing an upper limit, is obtained by calculating the field due to the pole face charge alone. This is a particularly useful procedure when the armature is neither long nor cylindrical. For this reason it is of value to have on hand the field reduction factors H/B_r for square and strip pole faces, given respectively by the functions

$$\frac{2}{\pi}\tan^{-1}\frac{b^2}{a(a^2+2b^2)^{1/2}} \quad \text{and} \quad \frac{2}{\pi}\tan^{-1}\frac{b}{a}, \tag{21}$$

where b is the face width and a is the gap width.



(a) (b)

Fig. 8—(a) Ceramic magnets in a cylindrical armature. (b) Ceramic magnets in a rectangular armature.

B. *Field of a Ceramic Magnet in a Rectangular Armature (special case)*

Thus far, we have considered magnet configurations which permit an exact field calculation to be made. An alternative approach, which forms the basis of magnetic-circuit engineering, is to make a guess at the shape of the magnetic field and then to apply the equations

$$\nabla \cdot \boldsymbol{B} = 0 \qquad \nabla \times \boldsymbol{H} = 0 \tag{22}$$

in order to determine the strength of the field. The first equation states that the magnetic induction is continuous, while the second equation states that the line integral of H around any closed path is zero.

The simplest case that may be analyzed is the case in which B is assumed to be uniform in the magnet and in the gap. This condition is met to a good approximation in the central field region of the circuit in Fig. 8(b), under the conditions $a, c \gg L, D$. Assuming that $H \simeq 0$ in the armature, we obtain the equations

$$H = B_r - H_m, \qquad DH = 2LH_m, \tag{23}$$

where H is the field strength in the gap and H_m is the field strength in the magnet. These equations have the solution

$$H = \frac{B_r}{1 + \dfrac{D}{2L}}. \tag{24}$$

The magnets operate at the point of maximum-energy product when $D = 2L$, that is the total magnet length is equal to the gap width. If, however, the magnet cross

section is fixed by the dimensions of the device, so that the magnet length is the only free variable, then different optimum dimensions are obtained. Considering the volume of the magnets plus the volume of the gap, the gap energy divided by this volume is given by

$$\frac{H^2Dac}{(D+2L)ac} = B_r^2 \frac{\left(2\frac{L}{D}\right)^2}{\left(1+\frac{2L}{D}\right)^3}. \qquad (25)$$

Thus, the useful magnetic energy per unit circuit volume is maximized when $L=D$, so that the total magnet length is equal to twice the gap width. This results in a field in the gap equal to $H=2(B_r/3)$, and a field in the magnet equal to $H_m=B_r/3$.

C. Iron Pole Faces

Ceramic pole pieces, as discussed in the previous sections, are quite effective in producing fields of about 2000 gauss along the magnetic axis. When we move away from the center of the field towards the gap edge we find that the field strength falls off gradually, starting about one gap width from the gap edge. At the intersection of the midplane between the pole faces and the gap edge the field strength will have dropped to about one-half of its value at the center. This "edge effect" may be calculated exactly for the case of two semi-infinite, uniform charge surfaces which are positioned relative to each other in the manner of condenser plates. These uniform charge surfaces represent the uniform pole distributions on the opposite faces of two semi-infinite, ceramic blocks. The field strength at the midplane is given then by

$$H = B_r\left(\frac{1}{2}+\frac{1}{\pi}\tan^{-1}\frac{2x}{a}\right), \qquad (26)$$

where a is the gap width and x is the horizontal distance between the gap edge and the field point, measured normal to the gap edge. This function is plotted in Fig. 9.

For applications requiring more uniform fields, the uniformity may be greatly improved by covering the ceramic pole faces with a thin sheet of iron. This frees the poles on the ceramic face and allows them to bunch up around the edges. In this way the field strength is raised near the edges and reduced near the center. The effectiveness of the pole redistribution in increasing field uniformity may be evaluated by the methods described in Section II-A. For this case we use the Schwarz transformation.[3]

$$dZ = \frac{(Z_1-1)(Z_1+1)}{Z_1}dZ_1. \qquad (27)$$

[3] Panofsky and Phillips, *op. cit.*, ch. 4.

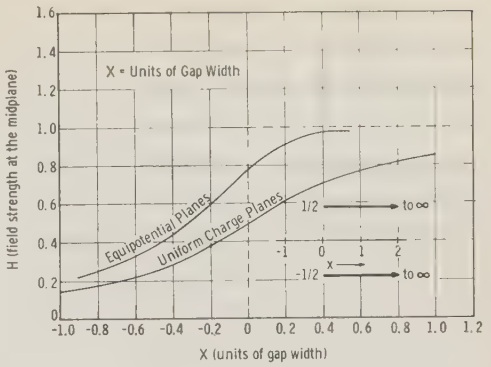


Fig. 9—Field strength near the edge of a magnetic gap.

This transformation bends the real axis in the Z_1 plane through an angle of $-\pi$ at $Z_1=\pm 1$ and through an angle of π at $Z_1=0$. In this way the real axis of the Z_1 plane becomes transformed into two semi-infinite, parallel lines in the Z plane, representing semi-infinite, parallel plates in two dimensions. The origin $Z_1=0$ is transformed to a point at infinity between the two plates, so that radial and azimuthal lines in the Z_1 plane represent lines of constant potential and constant stream in the Z plane. In this way the shape and strength of the field is determined for two semi-infinite, equipotential planes. The strength of the field at the midplane in the neighborhood of the gap edge is shown in Fig. 9. The addition of the iron pole faces produces a remarkable increase in field uniformity in comparison with the field due to ceramic pole faces. The fall-off in field strength near the edge does not begin until we are within 0.2 gap widths of the edge, and then on the midplane at the gap edge the field has dropped only to $\sim$0.8 of the central value.

Another aspect of the field related to its uniformity near the gap edge is its radius of curvature. The ratio of the gap width to the radius of curvature at the midplane is calculated in Fig. 10 for ceramic and iron pole faces. Again we see the effectiveness of the iron faces in increasing the field uniformity. The radius of curvature at the edge is on the order of the gap width.

The price of this increased field uniformity is a reduction in the useful energy contained in the gap. This arises because the total flux, including leakage, remains fixed. Then the bunching of the poles at the edges of the iron faces increases the amount of leakage flux at the expense of the flux inside the gap. An evaluation of this effect involves calculating the total energy in the gap for each case, and then comparing the two cases under the constraint that the total flux, including leakage, remains constant. This is a fairly difficult calculation which is easy to bypass with an experimental result. In Fig. 11 we show the measured field strength at the midplane between two ceramic pole pieces in a cylindrical armature. Although this is a three-dimensional rather

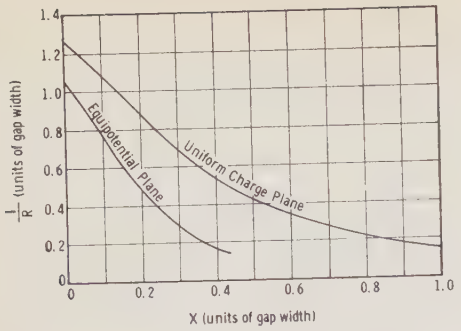


Fig. 10—Field curvature near the edge of a magnetic gap.

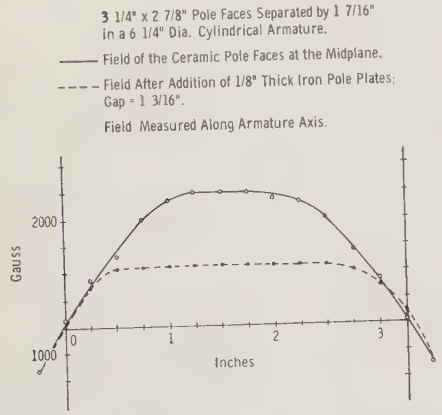

3 1/4" x 2 7/8" Pole Faces Separated by 1 7/16" in a 6 1/4" Dia. Cylindrical Armature.
Field of the Ceramic Pole Faces at the Midplane.
Field After Addition of 1/8" Thick Iron Pole Plates; Gap = 1 3/16".
Field Measured Along Armature Axis.
Gauss
2000
1000
0
1
2
3
Inches

Fig. 11—Measured field strength in the plane of a magnetic gap.

than a two-dimensional configuration, we see that the two-dimensional calculations are quite successful in predicting the behavior of the field in the neighborhood of the gap edge. We note also that in spite of the fact that the addition of the iron pole faces has reduced the gap width, the field strength is substantially lower than that obtained with the bare ceramic pole faces. Thus, the addition of the iron pole faces causes a very substantial reduction in the useful energy contained in the gap.

It is not easy to calculate the field strength in the gap when the ceramic poles are faced with iron. However, the following approximate procedure has met with some success. We may consider the pole density as being almost uniform except at the gap edges where the density becomes infinite. Calculations indicate that over a considerable range of geometry the plates of non-uniform pole density may be replaced by fictitious plates of uniform pole density which extend beyond the real plates by an amount equal to ~0.4 gap widths. This tends to preserve the pole density at the center and the total flux. The equivalent uniform pole density is given by

$$M = \frac{B_r}{4\pi} \cdot \frac{A}{A + 0.4Cd}, \tag{28}$$

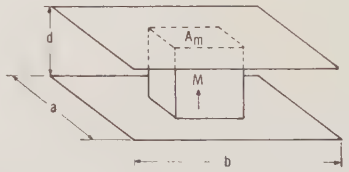
where A is the area of the real pole faces, C is the circumference of the pole face, and d is the gap width. With this value of M the field at the center is calculated for a uniform pole sheet which extends beyond the real faces by an amount $0.4d$. Once having the field at the center, we have it approximately everywhere by virtue of Fig. 9.

D. Field of a Magnet Between Two Rectangular Iron Plates

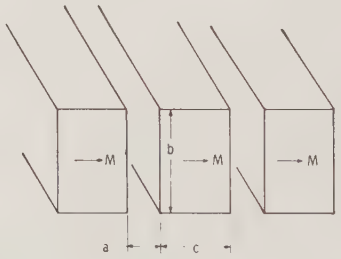
The results of the preceding section may be applied to the circuit of Fig. 12(a). We assume that the plates extend, in effect, a distance of about $0.4d$ beyond each edge. The field strength is accordingly given by (15) and (28).

$$H \simeq B_r \cdot \frac{A_m}{ab + 0.8(a+b)d} \cdot \frac{2}{\pi} \tan^{-1} \frac{a_1 b_1}{d\sqrt{d^2 + a_1^2 + b_1^2}}, \tag{29}$$

where A_m is the pole face area of the magnet, and $a_1 = a + 0.8d$, $b_1 = b + 0.8d$.



(a)



(b)

Fig. 12—(a) Magnet between iron planes. (b) A linear array of magnets.

E. Periodic Structures

The magnet for a high-speed Vaclon® pump consists of a linear array of long rectangular ceramic bars, as shown in Fig. 12(b), magnetized to provide a high magnetic field between the bars. For an infinite array, the field at the center of a gap is given by

$$H = \frac{2B_r}{\pi} \sum_{n=0}^{\infty} \left(\tan^{-1} \frac{b}{(2n+1)a + 2nc} - \tan^{-1} \frac{b}{(2n+1)a + 2(n+1)c} \right). \tag{30}$$

The nth term in this summation represents the contribution to the field by a magnet bar whose nearest face is

a distance of $(2n+1)(a/2)+nc$ from the field point. It is of some interest to show that this summation reduces in the limit $b \gg a$, c to a simple circuit formula. For $b \gg c$, the term difference in the parenthesis reduces to

$$\frac{2\dfrac{c}{b}}{1+\left[\dfrac{(2n+1)a+2nc}{b}\right]^2}. \tag{31}$$

If we let

$$x = \frac{(2n+1)a+2nc}{b}$$

then for a unit change in n, the corresponding change Δx in x is given by

$$1 = \frac{b}{2(a+c)}\Delta x. \tag{32}$$

Consequently, the summation reduces to

$$H = \frac{2B_r}{\pi}\cdot\frac{c}{a+c}\int_0^\infty \frac{dx}{1+x^2} = B_r\frac{c}{a+c}. \tag{33}$$

Allowing for the change in notation we see that this is the same equation used in (24) for ceramic poles in a rectangular armature.

As before, the gap energy per unit circuit length may be maximized. This yields the same result, *i.e.*,

$$c = 2a.$$

Thus, the total magnet length per gap is equal to twice the gap width. This optimization calculation has been carried out numerically with (30) for $b/a = 3$. This yields again the result $c \simeq 2a$, and we conclude that the optimum dimensions are not very sensitive to the value of b. For low values of b/a, such as $b/a = 3$, the field strength must be calculated by summing up the individual terms in (30). The series converges fairly rapidly for dimensions near optimum although, in general, it is surprising how slowly the field from distant gaps falls off in two dimensions.

V. The Combination of Ceramic and Steel Magnets

A. Comparison of Steel and Ceramic Magnets

The most commonly used steel magnetic material is Alnico V, which has a much higher remanence than Indox V ceramic, and a much lower field strength at the point of maximum BH product. This is shown in Fig. 2. The resulting value of maximum BH product for Alnico V is about 1.6 times greater than that for Indox V ceramic. This means, according to (10), that it requires a smaller volume of Alnico than of Indox in order to energize a given gap. In practice this may not work out due to the fairly high-flux leakage from the steel magnet itself.

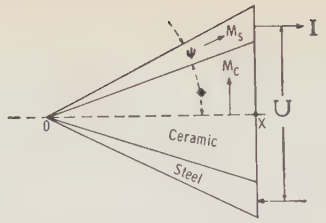
The consequence of the very high remanence of Alnico V is that only a small cross-sectional magnet area is required to obtain a high flux. On the other hand, the low coercive force results in a magnet element that is much longer than the gap spacing when high fields are required. Thus, steel magnets tend to be long and thin while ceramic magnets tend to be short and fat. The great virtue of a steel magnet is that the long, thin element can be cast into a horseshoe which surrounds the gap at a distance. A particular advantage of a ceramic magnet is that due to its very high coercive force it can be used directly in parallel with a high field gap.

B. The Wedge Magnet

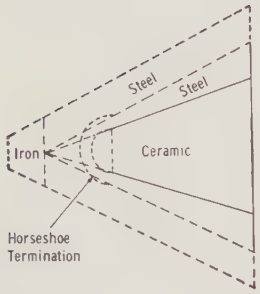
The main factor limiting the efficiency of steel horseshoe magnets is the energy contained in the leakage fields in the interior of the horseshoe. The strength of the leakage field is frequently on the order of 1000 gauss and it would provide therefore a good working field for a ceramic magnet whose magnetization is oriented antiparallel to this field. The induction of the ceramic magnet would then be added to that of the steel magnet so that a gap of larger area may be energized. We can look at this another way and say that the relatively high permeability of the steel magnet tends to confine to its body the fields of the ceramic magnet and to direct them into the gap. Considerations of this kind have been employed by Glass[4] in the design of steel magnets for electron beam focusing.

We will inquire now as to what is the optimum configuration for a ceramic magnet encased by a steel magnet. We wish that all volume elements of the structure operate at the point of maximum BH product. Further consideration reveals that in two dimensions this condition is met in the structure of Fig. 13(a) if the correct values are chosen for the angles θ and ψ. The steel magnet is magnetized in the radial direction while the ceramic magnet is magnetized in the azimuthal direction (although in practice it would probably be magnetized normal to the x axis). As in Section IV-B we will assume that we know the shape of the field. Outside of the steel magnet the H-field lines are assumed to form circles with a common origin at the left-hand corner of the magnet. In the upper half plane the H field inside of the steel magnet is assumed to be directed radially towards the origin while the H field in the ceramic magnet is in the negative azimuthal direction. The lines of induction B flow through the ceramic magnet in the positive azimuthal direction and out through the steel arms in the positive radial direction to a load (a gap) which is not shown. The circular H field outside of the magnet is a leakage field which may be neglected (as we will do) if θ is small compared with 90°.

[4] M. S. Glass, "Appraisal of permanent magnet materials for magnetic focusing of electron beams," *J. Appl. Phys.*, vol. 29, p. 403; March, 1958.



(a)



(b)

Fig. 13—(a) A combination ceramic-steel magnet. (b) Modifications.

Let B_c, H_c and B_s, H_s be the points of maximum energy product for the ceramic and steel magnets respectively. Then the angle θ is given by the equation $H_s x = H_c x\theta$ or

$$\theta = \frac{H_s}{H_c} \simeq \tfrac{1}{3} \text{ radian} \tag{35}$$

and ψ is given by the equation $B_c x = B_s x\psi$, or

$$\psi = \frac{B_c}{B_s} \simeq \tfrac{1}{6} \text{ radian}, \tag{36}$$

where the numerical values are given for the combination of Indox V and Alnico V. A wedge magnet of unit width and length x delivers a flux I and a potential U given by the equations

$$I = B_c x, \qquad U = 2H_s x. \tag{37}$$

This is a rather high-impedance output which may be suitable for electron beam focusing. The design may be modified for circular geometry in which the angle θ remains the same but the variation in thickness of the steel magnet is no longer linear with distance.

The impedance of a wedge magnet may be lowered, that is the flux may be increased, by adding a constant thickness of steel magnet to it. Also, the initial sharp corner may be avoided by starting out with a small horseshoe magnet as in Fig. 13(b).

VI. Conclusions

Measurements have been presented which show that the assumption of fixed magnetization is a useful approximation to the state of ceramic magnets. By means of this assumption a precise determination of remanence is made possible. Because of the fixed magnetization the field produced by isolated ceramic bars is easily computed, while the field due to ceramic magnets in an iron circuit may be calculated by image theory. It has been shown that by facing ceramic poles with a thin sheet of iron the field homogeneity is increased at the expense of reduced stored energy in the gap. A design optimization procedure has been found that violates in a specific case the ancient criterion that the magnet operate at the point of maximum BH product. An effort to make use of the leakage field from a steel horseshoe magnet has led to the design of a low-leakage ceramic-steel magnet in which both magnetic materials operate at their respective points of maximum BH product.

In applying the equations derived in this paper, the writer has been able to consistently predict field strengths within 10 per cent of their measured values. However, the ceramic-steel magnet has not been tested and so its actual performance must remain somewhat speculative.

Appendix

A. *Field of Two Equipotential Semi-Infinite Planes*

Integration of the Schwarz transformation

$$dZ = \frac{(Z_1 - a)(Z_1 + a)}{Z_1} dZ_1$$

yields

$$Z = \frac{Z_1^2}{2} - a^2 \ln Z_1 + C.$$

We want to evaluate C and a so that the real axis in the Z_1 plane transforms into two lines parallel to the x axis, which terminate on the y axis at points 0, $\pm i/2$. Setting Re $Z = 0$ and $Z_1 = \pm a$, gives

$$\text{Re } C = -\frac{a^2}{2} + a^2 \ln a.$$

Then

$$Z = \tfrac{1}{2}(Z_1^2 - a^2) - a^2 \ln \frac{Z_1}{a} + C'.$$

Then setting $Z_1 = a$, Im $Z = i/2$ and $Z_1 = -a$, Im $Z = -(i/2)$ gives, respectively,

$$C' = \frac{i}{2}, \qquad a^2 = \frac{1}{\pi}.$$

Since lines in Z_1 of constant angle are potential lines, and lines of constant radius are stream lines, we let $Z_1/a = \exp C''(V + iU)$, where U and V are the potential and stream functions. When $U = 0$ we move along the line $Z = x$, $i/2$ and when $C''U = \pi$, we move along the line $Z = x$, $-i/2$. For a potential difference of unity

we chose $C''=\pi$. Thus, the complete transformation is

$$Z=\frac{1}{2\pi}\left[e^{2\pi V}(\cos 2\pi U+i\sin 2\pi U)-1\right]-V-iU+\frac{i}{2}.$$

The field strength at the midplane may be obtained by setting $U=\frac{1}{2}$ and plotting $\partial V/\partial x$ against x. This is done by a graphical procedure involving the equations for $U=\frac{1}{2}$

$$-x=\frac{1}{2\pi}(e^{2\pi V}+1)+V, \qquad \frac{-dx}{dV}=e^{2\pi V}+1.$$

The charge density on the planes can be obtained by setting $U=0$ and calculating dV/dx from the equation of x against V

$$-x=\frac{1}{2\pi}(1-e^{2\pi V})+V.$$

It is clear that on the plane $U=0$, the condition $V=0$ gives $x=0$ so that the zero reference point for the stream function is at the edge of the plane.

Acknowledgment

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Electron Trigonometry—A New Tool for Electron-Optical Design*

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Summary—Electron-trigonometry is a new method to predict image formation in an electron-optical system, which may consist of one, or more, electron lenses of finite length, including electrostatic, magnetic or mixed lenses.

Electron-trigonometry by-passes trajectory tracing and the determination of cardinal points. Instead, it provides immediate information about size and position of an electron image.

To analyze an electrostatic system by "electron-trigonometry," a given axial potential distribution is approximated by segments, which are either voltage-linear, or parabolic. These segments are joined together with continuity of function and slope. Linear sections are replaced by equivalent drift tubes whose lengths are scaled appropriately. Parabolic sections are shown to observe certain basic triangle rules about the sum of three phase angles of flight, which are specifically defined for object, image, and lens volume. Magnification and position of electron images are immediately known from these same phase angles. This includes both real and virtual images, and lenses of either polarity.

A system including magnetic lenses is similarly treated. Such fields are subdivided into segments where the magnetic field is replaced by its average value while the axis potential is either linear or parabolic.

As a test case, electron-trigonometry is applied to the well-known system of a two-cylinder, bi-potential lens. This is a double lens system whose axis potential distribution is not simple. The results from electron-trigonometry are in excellent agreement with recorded data of performance.

In another application, electron-trigonometry is used to compute the *angular* magnification of a cathode ray tube with post-deflection acceleration. The outcome of this analysis is a novel type of scan-stabilizer. It uses an envelope with nonuniform spiral coating to realize a large, diverging lens. In this scan-stabilizer, beam propagation is essentially straight line. This results in a high value of deflection sensitivity (3.2 v/cm·kv) which does not change over a very wide range of screen voltages (from 2 to 20 kv).

I. Introduction

THE prediction of size and position of electron images is the basic problem of electron-optical design. Its solution hinges upon the theory of the thick electron lens which has been developed to its most general mathematical formulation many years ago [1].

Unfortunately, this theory is not of much help to the practicing engineer. It is derived from the paraxial equation which has successfully resisted solution in the great majority of cases. In order to determine the six cardinal points of the lens, one needs information about principal trajectories, in addition to a complete plot of equipotentials for the field. Considering the tedious job of step-by-step graphical integration, the engineer usually resorts to an automatic curve tracer, or winds up by preparing a program for a digital computer, while forgetting about the lens altogether.

This situation is unsatisfactory in at least one respect: Analysis by computer gives only one-shot answers for specific problems. By contrast, a good general theory can point out trends and functional relationships. Electron-trigonometry (ET) has been found to be a powerful tool for the design of electron-optical devices. This shall be exemplified in Section III.

The method of electron-trigonometry can also be used to analyze systems using magnetic as well as mixed electron lenses. This will be briefly outlined in Section IV.

II. Triangle Theorems in Electron-Optics

To apply electron-trigonometry, a given axis potential is first replaced by sections, whose voltage distribution is either linear, or parabolic. The linear sections make no contributions to lens action. They can always be replaced by equivalent drift spaces, whose potential is equal to one of the two boundary potentials of the linear section, but whose length is re-adjusted. The parabolic sections on the other hand, act as ideal, thick lenses.

The combination of one such lens section between two drift spaces is henceforth referred to as a "basic electron-optical system." For such systems, extremely simple triangle relations exist, which form the body of electron-trigonometry. Image size and position can then be followed through a more complex system by repeated application of these simple rules. No trajectory information is required.

Fig. 1 shows a basic system of electrostatic electron optics. It consists of an object space a, an image space b, and a lens section l. We assume both a and b to be *nonfocusing* spaces, *i.e.*, not contributing any radial restoring forces. Such nonfocusing spaces are either field free, or they are voltage-linear, both on axis, and also on the walls of a concentric cylinder. This condition can be realized in practice by coating a resistive spiral of *uniform* pitch on the inside of a cylindrical envelope. At the ends, plane metallic disks or meshes may be used to terminate the field, if necessary.

The focusing action in Fig. 1(a) takes place over a length l, which may fill part, or all, of the total system length. Within this volume, we assume an "ideal" lens

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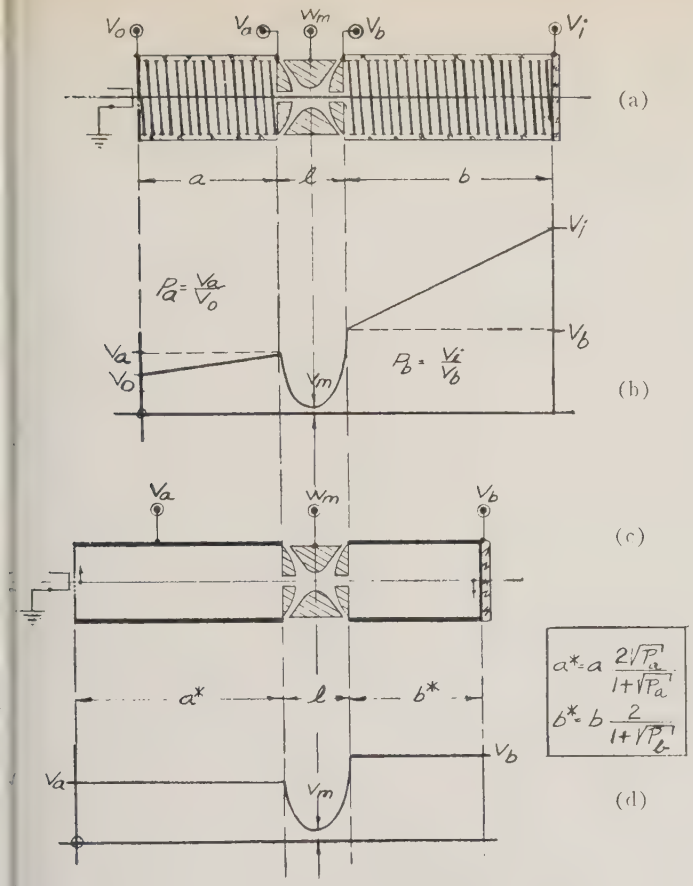


Fig. 1—Voltage-linear sections and equivalent drift spaces.

field, characterized by radial force components, that stay proportional to radius throughout, and are independent of axial position. This condition is met, if the potential distribution on axis is *parabolic*, *i.e.*, a general quadratic form in z. Rudenberg [2] has shown in 1948 that one way to realize such ideal lenses consists in using hyperbolic pole pieces similar to the arrangement shown in Fig. 1(a) and 1(c). Such structures are therefore also known as "hyperbolic lenses."

A. Linear and Constant Voltage Sections

The system of Fig. 1(a) includes two sections with linear acceleration, as indicated in Fig. 1(b) by the voltage ratios P_a and P_b for object- and image-space, respectively. This system can be replaced by an electron optical equivalent, as shown in Fig. 1(c). Here, the voltage-linear spiral sections a and b are replaced by field free drift spaces of a different length (a^* and b^*), meeting the lens section (l) at its original boundary voltages V_a and V_b. The new voltage distribution is shown in Fig. 1(d). Object and image sizes remain unchanged in the process, but the new drift-lengths are readjusted to meet (1):

$$a^* = a\frac{2\sqrt{P_a}}{1+\sqrt{P_a}} \qquad b^* = b\frac{2}{1+\sqrt{P_b}}\cdot \tag{1}$$

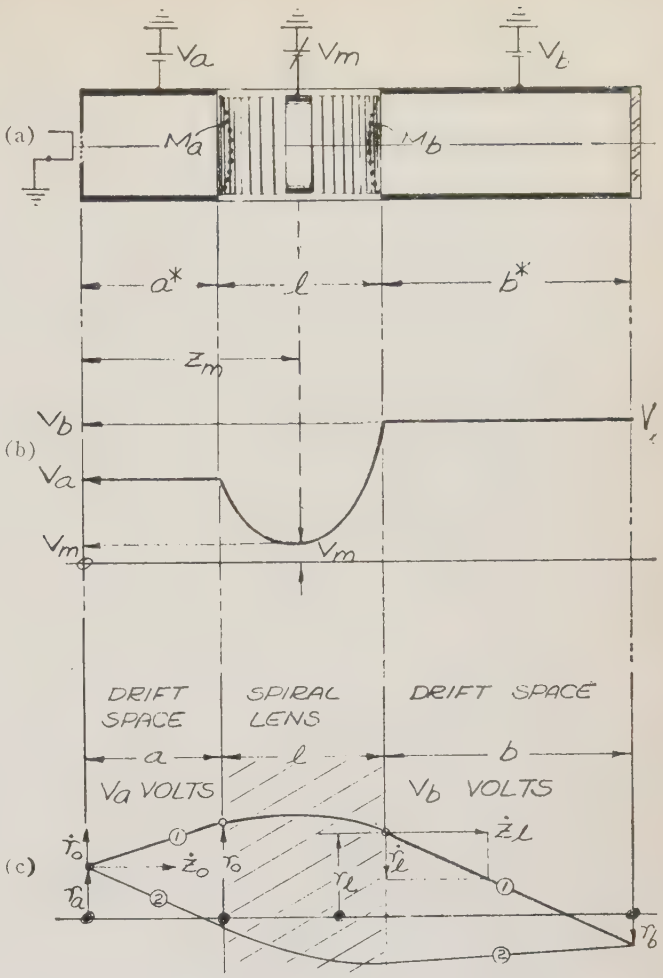


Fig. 2—Triangle theorem for positive electron lenses.

B. Hyperbolic Lens Section

The system is now transformed into that shown in Fig. 2(a) and 2(b). This arrangement is equivalent to Fig. 1(c) and 1(d), except that the hyperbolic lens is realized here in an alternative form. A lens cylinder of length l is coated on the inside by a set of *nonuniform* resistive spirals having variable, rather than constant, pitch. This "spiral lens" unit is equivalent to the embodiment used in Fig. 1, and either type may be used to provide the parabolic section of axis potential required for focusing. The lens cylinder has a much larger aperture, but it requires termination by a mesh (M_a, M_b) whenever a discontinuity of the gradient has to be realized.

Electron-optically, an ideal hyperbolic lens is fully characterized by its length l, and by its *power constant* c. The latter is the constant of proportionality between radial force per unit charge, and radial distance of axis. Hence, c is defined by

$$c = \frac{\frac{1}{e}\cdot F_r}{r}\cdot \tag{2}$$

On the other hand, F_r is linked to the space potential U by the relation

$$F_r = -e \cdot \frac{\partial U}{\partial r}. \tag{3}$$

In systems with cylinder-symmetry, we have the basic relation

$$U = V - \tfrac{1}{4}r^2V'' + \cdots, \tag{4}$$

where V is the axis potential. Thus, for a parabolic section,

$$\frac{\partial U}{\partial r} = -\frac{1}{2}V'' \cdot r. \tag{5}$$

Combining (2), (3) and (5) yields a general expression for the power constant of an electrostatic hyperbolic lens:

$$c = \frac{1}{2}\frac{d^2V}{dz^2}. \tag{6}$$

This may be positive or negative. In order for c to be a constant, the axis potential must be parabolic, *i.e.*, of the form

$$V = V_m + c(z - z_m)^2. \tag{7}$$

Here, V_m is the potential minimum or maximum and z_m is its axis coordinate [See Fig. 2(b)]. The distribution (7) on axis calls for a hyperbolic space potential U, which is found from (7) and (4):

$$U = V_m + c[(z - z_m)^2 - \tfrac{1}{2}r^2]. \tag{8}$$

This space potential U determines the design of the nonlinear spiral coatings and also the shape and curvature of any terminating meshes as used for instance in Fig. 2(a).

One important consequence of (8) is the fact that the potential distribution $W_{(z)}$ on the wall of a cylinder is parabolic, if this is true for its axis potential $V_{(z)}$. If D is the cylinder diameter, we have from (8),

$$W_{(z)} = V_{(z)} - \frac{c}{8}D^2. \tag{9}$$

An application for this will be described in Section III-C.

Finally, we can express the power constant c in terms of the voltages V_a and V_b at the lens boundaries z_a and z_b. Inserting this into (6) and (7) yields:

$$c = \frac{V_b - V_a}{l^2} - \frac{V_a'}{l}. \tag{10}$$

This simple relation for the power of a hyperbolic lens is best illustrated graphically. Fig. 3 shows part of some arbitrary functions of axial potential, featuring two sections A and B, which are practically straight, and an intermediate section C, which is curved. The latter is to be replaced by a parabola with continuity of slope at both ends. If this is done, the junction-

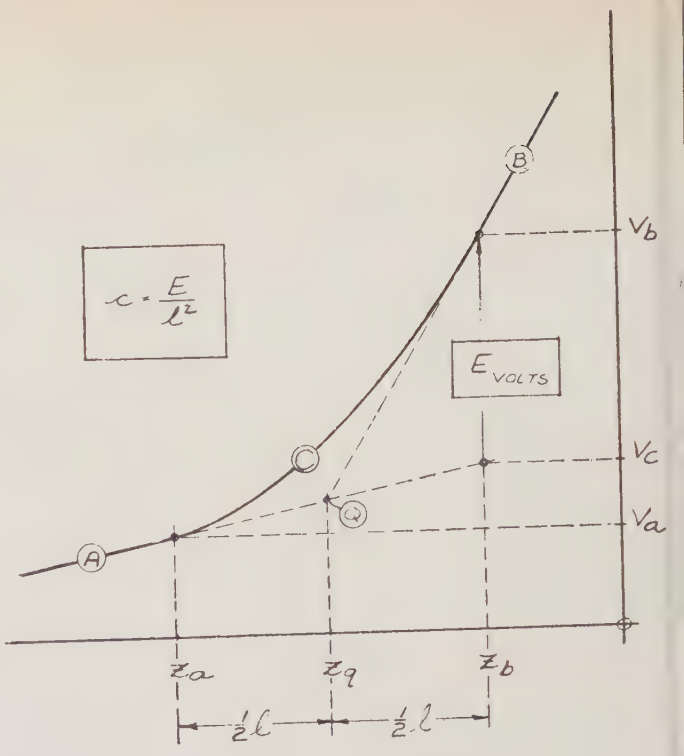


Fig. 3—Parabolic curve-fitting.

abscissas z_a and z_b will observe mirror symmetry with respect to the abscissa z_q of the intersection Q. As a result, the length l of the equivalent "Rudenberg" section is defined, if only one of the junction voltages, V_a or V_b is given. The power constant c of this parabolic lens section can then be read directly off the sketch below. We have

$$c = \frac{E}{l^2} \tag{10a}$$

where E is the voltage difference $V_b - V_c$, to be read off the graph as shown, both by magnitude and sign.

The power factor c is positive for a converging, and negative for a diverging, electron lens. Numerically, it ranges in practical devices from about 500 v/in² to about 50,000 v/in² and beyond. The smaller figure occurs in lens fields of large volume, such as the bulb spiral lens shown in Fig. 9. The high values for c are found in focusing lenses for cathode ray tubes, including the main focus lens in picture tubes. Still higher lens powers are used in the electron microscope and also in modern micro-gun designs, as used in cathode ray tubes for ultra-high resolution.

C. Triangle Theorem for a Positive System

We are now going to derive certain rules of electron-trigonometry, also called "triangle-theorems," or "delta rules" for short. These delta-rules are valid for electron optical systems such as Figs. 2 and 4, comprising one hyperbolic lens between two drift spaces, all of arbitrary length. For simplicity, we call such a basic system positive or negative, depending on the sign of the lens-power used.

Within the hyperbolic field, the electron trajectories are expressed by fairly simple equations (see appendixes). Thus, it is possible to describe the electron path through a basic system such as Fig. 2(c) by a set of four equations:

$$r_0 = r_a + a\frac{\dot{r}_0}{\dot{z}_0}, \tag{11}$$

$$r_l = r_0 \cos\varphi_c + \frac{\dot{r}_0}{\omega}\sin\varphi_c, \tag{12}$$

$$\dot{r}_l = \dot{r}_0 \cos\varphi_c - \omega r_0 \sin\varphi_c, \tag{13}$$

$$r_b = r_l + \dot{r}_l\frac{b}{\dot{z}_l}\cdot \tag{14}$$

In this set, the first and last equation simply describe linear propagation in object- and image-space, respectively. Eqs. (12) and (13) connect the radial position and velocity at the lens output to their counterparts at the lens input. Specifically, the parameters ω and φ_c are the "frequency" and the transit angle, or "phase," of a hyperbolic lens. These data are found from length l and power constant c of the lens, by using (15) and (16) (see Appendix I).

$$\omega = \sqrt{\eta c}, \quad \text{where} \quad \eta = e/m \tag{15}$$

$$\varphi_c = \sqrt{2}\cdot\tanh^{-1}\frac{l\sqrt{c}}{\sqrt{V_a}+\sqrt{V_b}}\cdot \tag{16}$$

Eqs. (11) through (14) can now be used to find the four unknowns including r_0, r_l, $\dot{r}_l$, and r_b. Specifically, the image size r_b then appears in the general form

$$r_b = M\cdot r_a + N\cdot\dot{r}_0, \tag{17}$$

where M and N are expressions which contain only axial, but no radial, velocity components [see (18) and (19)]. We have:

$$M = \cos\varphi_c - \frac{b}{\dot{z}_l}\omega\sin\varphi_c \tag{18}$$

$$N = \left(\frac{b}{\dot{z}_l}+\frac{a}{\dot{z}_0}\right)\cdot\cos\varphi_c + \left(\frac{1}{\omega}-\frac{a}{\dot{z}_0}\cdot\frac{b}{\dot{z}_l}\omega\right)\sin\varphi_c. \tag{19}$$

Inspection of (17) shows that r_b may become an "image" of r_a. In order for this to occur, it is necessary for the factor N to disappear from (17) (focusing condition). Once N is zero, the coefficient M in (17) indicates the magnification of the electron-image.

The focusing condition $N=0$ (19) assumes a surprisingly simple form if we introduce two new phase angles, one for the object- and one for the image-space. These are defined as

$$\text{object phase: } \varphi_a = \tan^{-1}\frac{a}{\sqrt{V_a}}\sqrt{\frac{c}{2}}, \tag{20a}$$

$$\text{image phase: } \varphi_b = \tan^{-1}\frac{b}{\sqrt{V_b}}\sqrt{\frac{c}{2}}\cdot \tag{20b}$$

In terms of these three phase angles, which include the lens phase from (16), image formation in a system such as Fig. 2(a), or its equivalent Fig. 1(c), can be expressed by a simple triangle relation:

$$\boxed{\varphi_a + \varphi_b + \varphi_c = \pi}\cdot \tag{21}$$

This permits one to find the image phase angle φ_b, and thus the image position b (20b), once object and lens phase are known, and vice versa. The image size is then also immediately known by introducing (20), (21) and (16) into expression (18) for M. For the system magnification this yields the simple relation

$$M_+ = -\frac{\cos\varphi_a}{\cos\varphi_b}\cdot \tag{22}$$

The above developments may be summarized as follows:

Triangle theorem for a positive system: If an electron image is formed by a system including one positive lens element between two drift spaces, the sum of the three phase angles for object-, lens- and image-space amounts to 180°. If lens phase plus object phase add up to more than 90°, the image is real and inverted. If that sum is less than 90°, the image is virtual and erect.

Applications for this triangle theorem will be given below. However, one interesting consequence of it may be mentioned here. By the use of (20a) and (20b), expression (22) for the magnification of a positive system is readily transformed into the form

$$M_+ = -\sqrt{\frac{\frac{2}{c}+\frac{b^2}{V_b}}{\frac{2}{c}+\frac{a^2}{V_a}}}\cdot \tag{22a}$$

For the lens power c, we know from (10) that it increases with decreasing thickness l of the lens. Thus, it follows from inspection of (22a) that the magnification of a positive system with one thick lens is intermediate between the "classical" value

$$M_+ = -\frac{b}{a}\sqrt{\frac{V_a}{V_b}}$$

for a very short lens, and the value $M_+ = -1$ which results from (22a) for a long lens, occupying most, if not all, of the total system length.

D. Triangle Theorem for a Negative System

A similar, but slightly different, delta-rule governs image formation in a system comprising one negative lens element between two drift spaces. A "negative system" of this kind is shown in Fig. 4(a), and its axis potential V in Fig. 4(b). As before, we introduce three

phase angles ϕ_a, ϕ_b, and ϕ_c for the two drift spaces, and for the lens volume, respectively. These are (see Appendix II for derivation)

$$\phi_a = \tanh^{-1} \frac{a}{\sqrt{V_a}} \sqrt{\frac{c_-}{2}}, \tag{23a}$$

$$\phi_b = \tanh^{-1} \frac{b}{\sqrt{V_b}} \sqrt{\frac{c_-}{2}}, \tag{23b}$$

$$\phi_c = \sqrt{2} \tanh^{-1} \frac{l\sqrt{c_-}}{\sqrt{V_a} + \sqrt{V_b}}. \tag{23c}$$

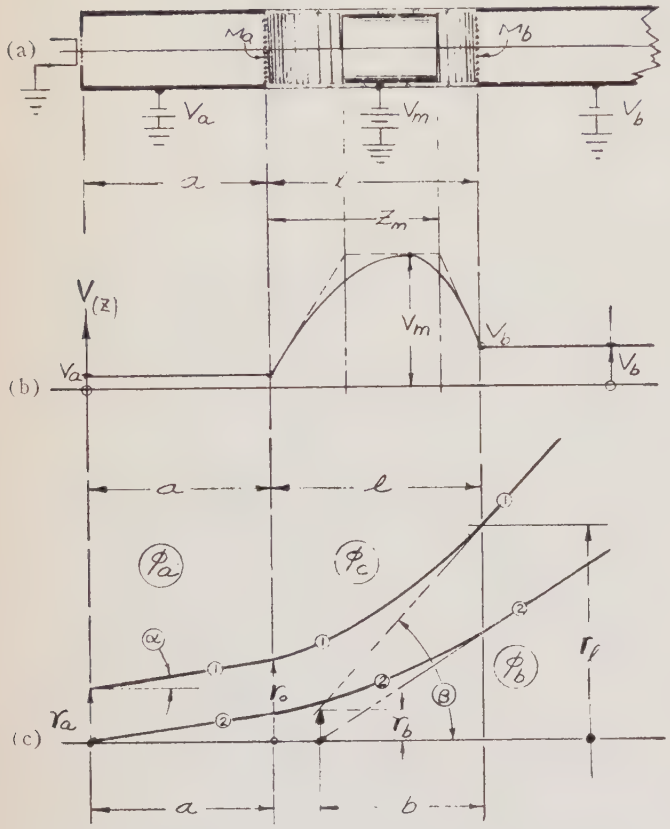


Fig. 4—Triangle theorem for negative electron lens.

Here, c_- is the absolute value of the power constant c of the negative lens: $c_- = |c|$. c_- is always positive, although c is negative in this case. A comparison with (16) and (20) for phases in a positive system shows that hyperbolic functions there have been replaced by cyclometric functions here and vice versa. In terms of these new phase angles ϕ, we find another triangle theorem to hold for a negative system:

Delta-rule for a negative system:

a) Image condition:

$$\boxed{\phi_a + \phi_b + \phi_c = 0}. \tag{24}$$

b) Magnification:

$$M_- = + \frac{\cosh \phi_a}{\cosh \phi_b}. \tag{25}$$

It can be seen from (25) that the negative electron lens is always noninverting, since M_- is positive. If the lens forms a virtual image, r_b, from a real object, r_a (see Fig. 4), this image is demagnified ($\phi_a > 0; |\phi_b| > \phi_a; M_- < 1$). By contrast, a magnified ($M_- > 1$) real image ($\phi_b > 0$) may be formed by a negative lens, if a virtual object is placed behind it. ($\phi_a < 0; \phi_b < |\phi_a|$). This situation exists for instance in electrostatic focusing by means of a bipotential lens (see Section III-A).

The two triangle-rules as introduced above are summarized for practical use in Fig. 5. For a negative electron lens, the delta theorem is stated in two versions which are supplementary. The alternate form for ϕ_a^* or ϕ_b^* is needed, if the expressions for ϕ_a or ϕ_b exceed unity, *i.e.*, the range of the function $\tanh^{-1}\phi$.

In any practical application of electron-trigonometry, a rule of signs should be observed as follows: trace a path from object, through lens, to image. Any vector (a, b, l) which points in the direction of electron flow,

Converging lens c positive; $c_+ = \|c\|$	Diverging lens c negative; $c_- = \|c\|$	
$\varphi_a = \tan^{-1} \frac{a}{\sqrt{V_a}} \sqrt{\frac{c_+}{2}}$	$\phi_a = \tanh^{-1} \frac{a}{\sqrt{V_a}} \sqrt{\frac{c_-}{2}}$	$\phi_a^* = \tanh^{-1} \frac{1}{a} \sqrt{\frac{2V_a}{c_-}}$
$\varphi_b = \tan^{-1} \frac{b}{\sqrt{V_b}} \sqrt{\frac{c_+}{2}}$	$\phi_b = \tanh^{-1} \frac{b}{\sqrt{V_b}} \sqrt{\frac{c_-}{2}}$	$\phi_b^* = \tanh^{-1} \frac{1}{b} \sqrt{\frac{2V_b}{c_-}}$
$\varphi_c = \sqrt{2} \cdot \tanh^{-1} \frac{l\sqrt{c_+}}{\sqrt{V_a} + \sqrt{V_b}}$	$\phi_c = \sqrt{2} \cdot \tan^{-1} \frac{l\sqrt{c_-}}{\sqrt{V_a} + \sqrt{V_b}}$	$\phi = \sqrt{2} \cdot \tan^{-1} \frac{l\sqrt{c_-}}{\sqrt{V_a} + \sqrt{V_b}}$
$\varphi_a + \varphi_b + \varphi_c = \pi$	$\phi_a + \phi_b + \phi_c = 0$	$\phi_a^* + \phi_b^* + \phi_c = 0$
$M_+ = - \frac{\cos \varphi_a}{\cos \varphi_b}$	$M_- = + \frac{\cosh \phi_a}{\cosh \phi_b}$	$M_- = - \frac{\sinh \phi_a^*}{\sinh \phi_b^*}$

Fig. 5—Rules of electron-trigonometry.

counts positive; if opposed to the flow, negative. The same signs also apply to the respective phase angles used in the triangulations.

III. Applications of Electron-Trigonometry

To illustrate the use of electron-trigonometry, two applications of it are presented below. The first is of the nature of a test case: it is the well-known problem of the two cylinder lens [3]. This is a two-lens system, which will be analyzed by means of the delta-rules for both positive and negative lens elements.

The second application given here shows how electron-trigonometry may be used to compute angular, rather than lateral, magnification. Here, the theory points the way to a novel type of post deflection accelerator. This case is presented to show the use of the new method for the design of electron-optical devices.

A. The Two-Cylinder, Bi-Potential Lens

The system under consideration is shown in Fig. 6(a). Two cylinders of equal diameter, and of length l_1 and l_2, are connected to voltages V_1 and V_2, respectively. The grounded cathode K illuminates a wire mesh at the entrance to the first cylinder. We wish to find size and position of the electron image of this mesh, as a function of the voltage ratio $P = V_2 : V_1$. For two cylinders of equal diameter D and narrow spacing, as shown in Fig. 6(a), F. Gray [4] has given an accurate expression for the axis potential:

$$V = \frac{V_1 + V_2}{2} + \frac{V_2 - V_1}{2} \cdot \tanh\left(2.64\,\frac{z}{D}\right). \quad (26)$$

We are replacing this function by two parabolic segments which cover the same range of voltages (from V_1 to V_2), and which are joined at the same slope as the original function (26). This approximation is shown in Fig. 6(b). The two parabolas are

$$1) V = V_1 + c_+(z + \lambda)^2, \quad (27)$$

$$2) V = V_2 + c_-(z - \lambda)^2. \quad (28)$$

The first represents a positive lens with the power constant

$$c_+ = \frac{V_2 - V_1}{2\lambda^2}. \quad (29)$$

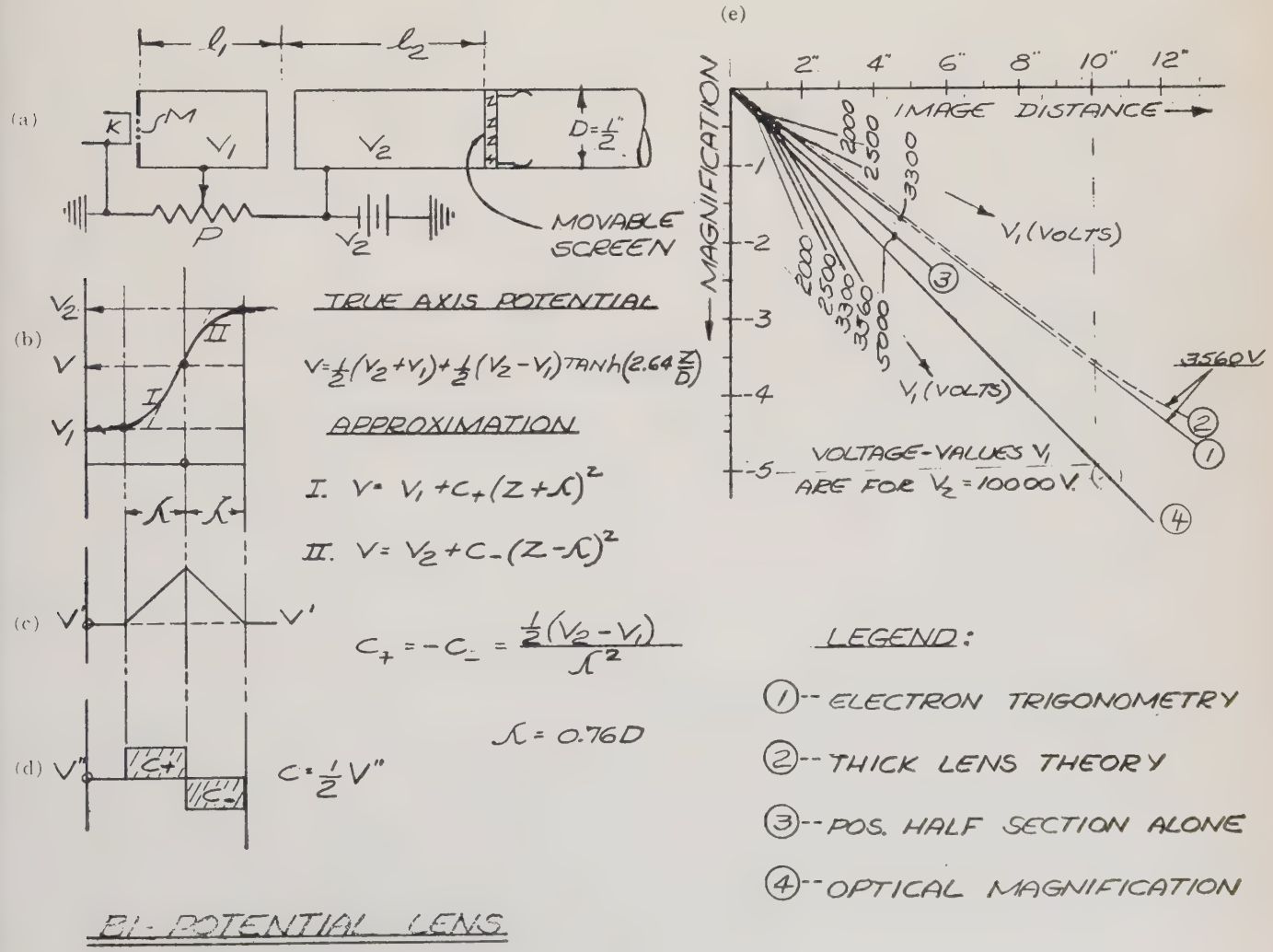


Fig. 6—Analysis of the bi-potential lens.

The second is a negative lens with the same power:

$$c_- = -c_+. \tag{30}$$

The thickness of both lenses follows from (26),

$$\lambda = 0.76D = 0.38''. \tag{31}$$

Both lens sections are shown in Fig. 6(d). We apply ET to the first lens. The first cylinder has the dimensions $l_1=2$ in, $D=\frac{1}{2}$ in. The object distance then becomes, with (31): $a=l_1-\lambda=1.62$ in. The object phase angle then follows with (29), (31), and $P=V_2:V_1$ as:

$$\varphi_a = \tan^{-1} 2.14\sqrt{P-1}. \tag{32}$$

The lens phase follows from (16) as:

$$\varphi_c = \sqrt{2}\tanh^{-1}\frac{\sqrt{P-1}}{\sqrt{2}+\sqrt{P+1}}. \tag{33}$$

The image phase now follows from the triangle rule for a positive lens,

$$\varphi_b = \pi - \varphi_a - \varphi_c. \tag{21}$$

Hence, the image distance b of the first lens is [from (20b)]

$$b_+ = \sqrt{2}\cdot\lambda\frac{\sqrt{P+1}}{\sqrt{P-1}}\cdot\tan\varphi_b, \tag{34}$$

and the magnification of the first lens [from (22)]

$$M_+ = -\frac{\cos\varphi_a}{\cos\varphi_b}. \tag{22}$$

This has been plotted in Fig. 6(e) as curve no. 3 which presents the action of the first, positive lens section acting alone.

We now proceed to the second, diverging lens section of the two-cylinder system. This negative lens section looks at the image defined by (34) and (22) as a virtual object, and projects a real, magnified image from it. Hence, we enter b_+ from (34) into the triangulation of the second lens as a negative object distance:

$$a_2 = -b_1. \tag{35}$$

Hence, the object phase becomes:

$$\phi_a = \tanh^{-1}\frac{\sqrt{2}\cdot\lambda}{a}\sqrt{\frac{P+1}{P-1}}. \tag{36}$$

The phase angle of the negative lens follows from (23c) as:

$$\phi_c = \sqrt{2}\cdot\tan^{-1}\frac{\sqrt{P-1}}{\sqrt{2P}+\sqrt{P+1}}. \tag{37}$$

The image phase follows from the triangle theorem for negative lenses (24),

$$\phi_b = -\phi_c - \phi_a. \tag{24}$$

Finally, the image distance is found from (23b),

$$b = \frac{2\lambda}{\tanh\phi_b}\sqrt{\frac{P}{P-1}}. \tag{38}$$

The magnification of the negative section is known from (25) of the second triangle theorem,

$$M_- = -\frac{\sinh\phi_a^*}{\sinh\phi_b^*}. \tag{25}$$

The size of the final image observed on the screen is then the product of the magnifications from both lenses, as found from (22) and (25),

$$M_{\text{total}} = M_+\cdot M_-. \tag{39}$$

This is plotted in Fig. 6(e) as curve no. 1. One can see, by comparison with curve no. 3, that the combined system requires a much lower voltage V_1 than the positive section alone to focus at a given throw. It is also interesting that the throw increases fast (from 4 in to 12 in) with only a 3 per cent change of voltage on the first cylinder (300 v out of 10,000 v on the screen). Also, the over-all magnification is somewhat smaller for the composite system than for the positive lens component alone, and both values are again smaller than the distance ratio, *i.e.*, the optical magnification by a thin lens, placed between the two cylinders.

It is noteworthy that the electronic magnification when plotted vs object distance, is essentially a straight line through the origin. This justifies the introduction of an "electron-optical factor" k,

$$k = m_{\text{electronic}}:m_{\text{optical}}. \tag{40}$$

This factor is a characteristic number for any electron-optical system. In accelerating systems of this type, k is smaller than unity, but not as small as the root-of-voltage formula, which is often used as a first approximation.

More important, however, than these details seems to be the experience that electron-trigonometry has here been applied to an electron-optical device which is definitely "non-ideal." The results are in excellent agreement with data gathered by conventional methods. Curve no. 2 in Fig. 6(e) presents the performance of this two-cylinder lens as recorded in test-book literature [3]. The agreement is within the accuracy of slide rule and of measurement.

B. Triangle Theory for Angular Magnification

In some applications of cathode ray tubes, one is interested in the stabilization of deflection in the presence of post deflection acceleration (PDA). In other applications, one wishes to magnify some small initial sweep by the action of a suitable electrostatic field after deflection. To design such systems, it is necessary to predict the amount of angular magnification which can be realized within a given system.

Fig. 7 shows the basic configuration for scan magnification. Point A is the center of a small, initial deflection α. After drifting through a field-free space a, the predeflected ray enters into a lens field of the length l. It then leaves this field under an increased angle β and traverses a second drift space, before reaching the screen. Intersection of the outgoing ray with the axis defines a point B, which is by definition an image point of A. Hence, its position can be found by electron-trigonometry, as shown before. The actual scan magnification $S = R_b/R_a$, as observed on the screen, can be read off Fig. 7 as

$$S = \frac{R_b}{R_a} = \frac{\tan \beta}{\tan \alpha}\left[1 - \frac{a + l + b}{L}\right]. \qquad (41)$$

Here, the second factor indicates the influence of drift spaces surrounding the lens. Note that b is the only variable in this term. It counts negatively for a diverging lens, as shown in Fig. 7. The first factor,

$$m_\alpha = \tan \beta / \tan \alpha, \qquad (42)$$

is an expression for angular magnification. A formal analysis of the electron path in Fig. 7, assuming the lens volume to be confined between plane boundaries, leads to an expression of the form

$$m_\alpha \cdot M \cdot \sqrt{P} = 1; \qquad P = V_b/V_a. \qquad (43)$$

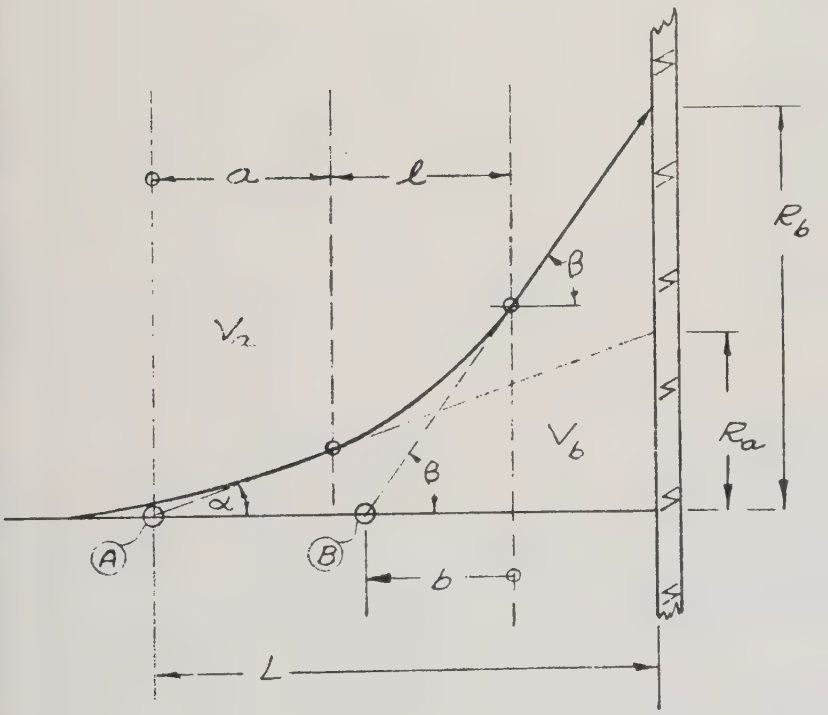
This equation provides the desired link between angular and lateral magnification. It closely resembles the fundamental theorem by Helmholtz-Lagrange, except for the fact that a ratio of sines, rather than tangents, occurs there [5]. Numerically, this difference stays within 10 per cent for sweep angles up to 50°.

The general expression for scan magnification is a combination of (41) and (43):

$$S = \frac{1}{M \cdot \sqrt{P}}\left[1 - \frac{a + l + b_{(P)}}{L}\right] \qquad (44)$$

This formulation suggests that:

1) Acceleration opposes, and deceleration enhances, scan magnification.



$$m_\alpha = \frac{\tan \beta}{\tan \alpha}$$

$$P = V_b : V_a$$

$$M_- = + \frac{\cosh \phi_a}{\cosh \phi_B}$$

$$M_+ = - \frac{\cos \phi_a}{\cos \phi_b}$$

SCAN MAGNIFICATION:

$$S = \frac{R_b}{R_a} = \frac{1}{M \cdot \sqrt{P}}\left[1 - \frac{a+l+b}{L}\right]$$

ANGULAR MAGNIFICATION:

$$m_\alpha \cdot M \cdot \sqrt{P} = 1$$

Fig. 7—Analysis of scan magnifiers.

2) A system using a diverging lens ($M<1$; b negative) is more promising than a system using a positive lens (b positive). By comparison, the system with a positive scan-magnifier lens requires greater length and is less efficient.

Reference is now made to Fig. 8(a). This shows one way to realize a large negative lens of the accelerating type. The lens is constructed by the deposition of a nonuniform, resistive spiral coating on the inside of a cylindrical envelope. This "spiral lens" must have a decreasing, *i.e.*, nonuniform, turn density in the direction toward the screen. The latter, in turn, is surrounded by a conductive band, which projects by $\frac{1}{4}\sqrt{2}$, or 0.35, screen diameters backward towards the gun. A drift space (a) after deflection is an essential component in this spiral-lens tube as we shall see. We compute, by electron-trigonometry, the scan magnification S in this tube using (44). Since the tube has no drift space after the lens l, we have $a+l=L$, and (44) reduces to

$$S = \frac{1}{M_- \cdot \sqrt{P}} \cdot \frac{-b}{L} \cdot \tag{45}$$

Here, M is the magnification of the negative spiral lens, given above as

$$M_- = \frac{\cosh \phi_a}{\cosh \phi_b} \cdot \tag{25}$$

We start by computing the object phase ϕ_a from (23a). Introducing the lens power,

$$c_- = \frac{V_b - V_a}{l^2}, \tag{46}$$

and the post-acceleration factor $P = V_b/V_a$, (23a) becomes

$$\phi_a = \tanh^{-1} 0.71 \cdot \frac{a}{l} \cdot \sqrt{P-1}. \tag{47}$$

Next, we formulate the lens phase ϕ_c. Its basic equation (23c) transforms with (46) into

$$\phi_c = \sqrt{2} \tan^{-1} \frac{\sqrt{P-1}}{1+\sqrt{P}} \cdot \tag{48}$$

From (47) and (48), we now get the image phase ϕ_b, by using the delta-theorem for negative lenses:

$$\phi_b = -\phi_a - \phi_c. \tag{24}$$

This phase angle is negative. It permits to find the lateral magnification M_- by going into (25) with ϕ_b, and with ϕ_a from (47). Knowing ϕ_b, we also find the image distance b. From (23b) we obtain, after some manipulation,

$$b = l \sqrt{\frac{2P}{P-1}} \cdot \tanh \phi_b. \tag{49}$$

We now have all necessary data to use (45) for the computation of scan magnification in the spiral-lens tube. This has been done in Fig. 8(c) for a wide range of post acceleration, and for various values of the funnel-to-bulb ratio a/L. It will be seen from Fig. 8(c) that in the absence of any object-drift length ($a=0$), the diverging spiral lens is completely ineffective! At a post acceleration of 10:1, the deflection loss then amounts to 50 per cent, which is no better than standard practice with a uniform spiral anode. However, as the relative funnel length a/L is increased, the spiral lens section l becomes shorter, and its power increases with $1/l^2$. The spiral thus becomes rapidly more effective in opposing deflection losses from post acceleration. The point of balance is reached for $a/L=0.56$. Beyond that, there is some slight overcompensation (about 20 per cent). At the critical funnel ratio, the theory predicts the raster size to stay constant within 3 per cent over a 10:1 range of post acceleration!

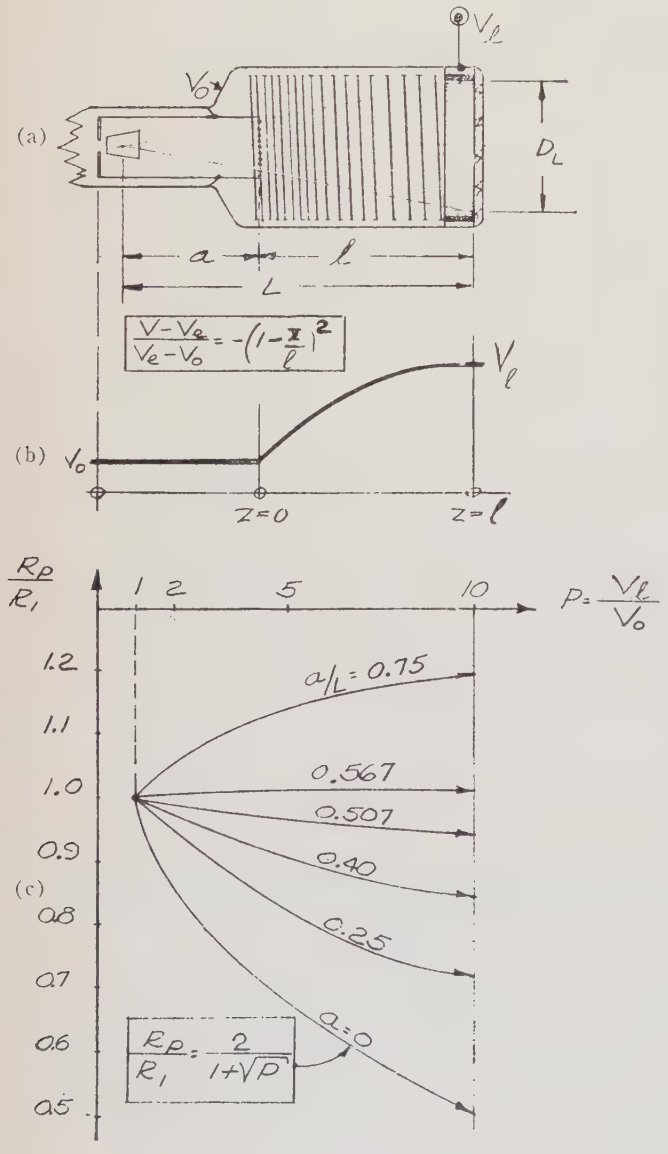


Fig. 8—Scan stabilization by a spiral lens.

C. An Experimental Test

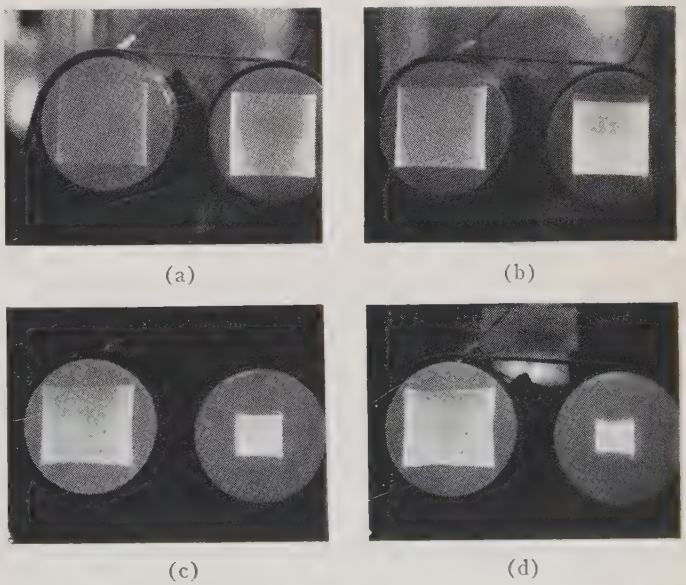
The above predictions of electron-trigonometry were put to a test by constructing an experimental cathode ray tube, which was built in close correlation to the theoretical design. Fig. 9 shows a photograph of this tube, whose purpose it was to achieve constant deflection sensitivity over a wide range of screen voltages.

The tube consists essentially of a cylindrical bulb, coated by a resistive spiral with nonuniform pitch. The specific turn-density per inch varies in such a manner that the voltage gradient decreases linearly with progress toward the screen. The screen is metallized, and framed by a conductive band, which projects back by 0.36 screen diameters. The voltage distribution along the wall is parabolic, as shown in Fig. 8(b). This establishes within the bulb a diverging lens, which approaches the ideal field quite well, as confirmed by experience.

A second, essential feature is the fairly long unipotential drift tube between the internal deflection yoke [6] and the spiral input. This funnel is at the average deflection potential, *i.e.*, the second anode voltage. At the far end, it is covered by a wire gauze.

Fig. 10 shows a photograph of the actual performance of this spiral lens stabilizer. Four conditions are shown with screen voltages going through 1–3.5–8.5 and 21 kv, while the second anode was held constant at 1000 v. The spiral lens stabilizer (left), is shown side-by-side with a conventional monoband intensifier (right), all voltages, including deflection and acceleration, being connected in parallel on both tubes. The spiral tube was found to have a high value of deflection sensitivity —3.2 v/cm. kv—and this value did not change over the very wide range of screen voltages used in the test.

Fig. 9—Spiral-lens tube assembly.



(a) (b) (c) (d)

Fig. 10—Scan stabilizer test: spiral-lens tube (left)—monoband tube (right). Second anode voltage: 1000 volts. Screens operating at: (a) 1000 v, (b) 3500 v, (c) 8500 v, (d) 21,000 v.

IV. Electron-Trigonometry for Magnetic and Mixed Systems

In the preceding sections, the method of electron-trigonometry has been presented in connection with purely electrostatic electron-optics. However, the method applies equally well to electron-optical systems which employ magnetic lenses, either alone, or in combination with electrostatic lenses. To prepare a problem of this kind for electron-trigonometry, the system is again subdivided into sections, which have a linear, or a parabolic, potential distribution. The magnetic field, on the other hand, is replaced by the average value which it has across each section [7].

Using polar coordinates r and θ, the following equations hold for each such section:

$$\ddot{r} - r\dot{\theta}^2 = -\eta r\left[\tfrac{1}{2}V'' + \dot{\theta}B\right], \tag{50}$$

$$\frac{1}{r}\cdot\frac{d}{dt}\left(r^2\dot{\theta}\right) = \eta\dot{r}B \tag{51}$$

$$\ddot{z} = \eta V'. \tag{52}$$

Undistorted image formation is one solution of these equations, since, for B = constant, (51) is satisfied by

$$\dot{\theta} = \tfrac{1}{2}\eta B = \text{constant}. \tag{53}$$

This implies a rotation of the entire image at the local Larmor frequency, independent of radius. It then follows from (50) that

$$\ddot{r} = -\eta r\left[\frac{1}{2}V'' + \frac{\eta B^2}{4}\right]. \tag{54}$$

Hence, an observer on a reference plane through axis, which rotates with the Larmor frequency, will find the beam subject to a composite lens field with an effective power constant of

$$c_{\text{total}} = c_m + c_e, \tag{55}$$

where

$$c_m = \tfrac{1}{4}\eta B^2 \tag{55a}$$

and

$$c_e = \tfrac{1}{2}V''. \tag{55b}$$

Here, the magnetic lens power c_m is always positive and proportional to e/m, while the electric lens power is independent of e/m, and can have either polarity. Hence, our observer on the rotating reference plane gets correct results from electron-trigonometry, if he uses (56) for the phase φ_c of the mixed lens,

$$\varphi_c = \sqrt{2}\tanh^{-1}\frac{l\sqrt{c_e + c_m}}{\sqrt{V_0} + \sqrt{V_l}}. \tag{56}$$

Meanwhile, the magnetic field rotates the reference plane through an angle of

$$\psi = \dot{\theta}\cdot T = \tfrac{1}{2}\eta B\cdot T, \tag{57}$$

where T is the transit time through the lens. If the axis potential has a quadratic form

$$V = a + bz + cz^2, \tag{58}$$

then it can be shown by simple integration that the transit time T is always given by:

$$T = \int_0^l \frac{dz}{\sqrt{2\eta V}} = \sqrt{\frac{2}{\eta c_e}}\cdot\tanh^{-1}\frac{l\sqrt{c_e}}{\sqrt{V_0} + \sqrt{V_l}}. \tag{59}$$

Here V_0 and V_l are the potentials at the two boundaries to the section. This leads to (60) for the image rotation per section:

$$\psi = \sqrt{2}\cdot\sqrt{\frac{c_m}{c_e}}\cdot\tanh^{-1}\frac{l\sqrt{c_e}}{\sqrt{V_0} + \sqrt{V_l}}. \tag{60}$$

A comparison of (56) for the lens phase φ_c, with (60) for the angle of rotation ψ, shows that the two angles are generally *not* equal, as long as the mixed lens has an electrostatic component ($c_e \neq 0$). This becomes even more evident, if elemental angles are written down for a very thin slice of the mixed field. Eq. (56) then reduces to

$$\Delta\varphi_c = \sqrt{\frac{c_e + c_m}{2V}}\cdot\Delta z \tag{56a}$$

and (60) becomes[1]

$$\Delta\psi = \sqrt{\frac{c_m}{2V}}\cdot\Delta z. \tag{60a}$$

It is now evident, that the two angles φ_c and ψ become identical, if the lens section is purely magnetic ($c_e = 0$). This includes the practically important case of magnetic focusing in the presence of a constant, or zero, axial electric field ($V = a + bz$). In systems of this kind the phase concept assumes a physical meaning: it can be directly observed as the angle of rotation of an electron image!

V. Automatic Computation

During the past two years electron-trigonometry has been applied in these laboratories to a great variety of problems in electron-optics. Quite generally the method has been found to provide reliable, and often quantitative, guidance during the early design stages. In view of this experience and in order to save time, it was decided to prepare a program of electron-trigonometry for an IBM 709 digital computer. This work was done under the direction of B. A. Findeisen of this Department, and in cooperation with the Heavy Military Equipment Department of the General Electric Company. At present we are in a position to compute the over-all electron-optical effect of as many as 250 subsections of a given axis potential. In this program, the computer automatically evaluates the power constant of each subsection and then uses this data to find the new image information by "electron-trigonometry." Besides saving time, this method offers the important advantage of eliminating the factor of human judgment, which is ever-present, if field-approximations have to be found by graphical methods.

Appendix I

Phase Angle in a Positive Lens

Within a hyperbolic lens, the power factor c is a constant, and the equations of motion are linear and harmonic. They read

$$\ddot{r} = -\eta c r, \tag{61}$$

$$\ddot{z} = 2\eta c(z - z_m), \tag{62}$$

[1] Eq. (60a) leads to the well-known expression for image rotation by a thick magnetic lens if the value for c_m from (55a) is introduced prior to integration of (60a) through the lens field.

where z_m is the position of the minimum axis potential, and $\eta = e/m$. At the start, we have

$$t = 0 \quad z = 0 \quad r = r_0 \quad \dot{z}_0 = \sqrt{2\eta V_0} \tag{63}$$

and at the end,

$$t = T \quad z = l \quad r = r_l \quad \dot{z}_e = \sqrt{2\eta V_l}. \tag{64}$$

The general solution of (61) and (62) is

$$r = r_0 \cos \omega t + \frac{\dot{r}_0}{\omega} \sin \omega t \tag{65}$$

$$z = z_m(1 - \cosh \sqrt{2}\omega t) + \frac{\dot{z}_0}{\sqrt{2}\omega} \sinh (\sqrt{2}\omega t). \tag{66}$$

Here, we introduced

$$\omega = \sqrt{\eta c} \tag{15}$$

as the frequency of the lens. Fitting this into the boundary condition (64) yields

$$l = z_m[1 - \cosh \sqrt{2}\varphi_c] + \frac{\dot{z}_0}{\sqrt{2} \cdot \omega} \sinh (\sqrt{2}\varphi_c) \tag{67}$$

and

$$\dot{z}_e = \dot{z}_0 \cosh (\sqrt{2}\varphi_c) - \sqrt{2}\omega z_m \cdot \sinh (\sqrt{2}\varphi_c). \tag{68}$$

These are two equations for the position z_m of the potential minimum, and also for the lens phase φ_c. Solving for the latter, we obtain the expression

$$\varphi_c = \omega T = \sqrt{2} \cdot \tanh^{-1} \frac{l\sqrt{c}}{\sqrt{V_0} + \sqrt{V_l}} \tag{16}$$

as used in the text.

Appendix II

Phase Angle in a Negative Lens

For a negative power constant, $c < 0$ $|c| = c_-$, the lens frequency $\omega = \sqrt{\eta c}$ becomes imaginary. Introducing, instead, its absolute value

$$\Omega = \sqrt{\eta \cdot c_-} \tag{69}$$

the equations of motion now read:

$$\ddot{r} = \Omega^2 r \tag{70}$$

$$\ddot{z} = -2\Omega^2(z - z_m). \tag{71}$$

Their solutions are

$$r = r_0 \cosh (\Omega t) + \frac{\dot{r}_0}{\Omega} \sinh (\Omega t), \tag{72}$$

$$z = z_m(1 - \cos (\sqrt{2}\Omega t)) + \sqrt{\frac{V_0}{c_-}} \sin (\sqrt{2}\Omega t). \tag{73}$$

Eq. (23c) for the lens phase ϕ_c follows from fitting (73) into boundary conditions similar to (63) and (64) in Appendix I.

The expressions (23a) and (23b) for object and image phase of the negative lens follow from a system analysis, similar to the one given in the text for a system including a single, positive, electron lens.

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New FM-AM Method of Compression, Expansion, and Multiplication*

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Summary—**A new method of employing the combination of two signals has been developed for magnitude compression or expansion of audio or video signals and for multiplication or other uses involving modulation with separated components (*e.g.*, instrumentation and remote control). The first signal frequency-modulates a local oscillator, whose output is limited and then amplitude-modulated by the second signal; the combined FM-AM signal is detected in such a way that the first signal is restored to its original form, but modified in amplitude by the second signal. In compression-expansion, the method controls the signal level with great speed and over a wide band without introducing noise, "bounce," or dc shift into the output; no delicate balancing or matched-tube circuits are required. In multiplication, the product of two voltages can be formed by means of a simple circuit that is extremely suitable for analog computers and similar applications.**

Introduction

THE method described below was initially developed for the purpose of achieving nearly instantaneous audio compression and expansion that would be free from the disadvantages of other methods in current use. Apart from the description contained in U. S. patents issued[1] and pending (covering, among others, all of the uses described in the present paper), the first public report[2] of the new method was largely concerned with the audio application. The present paper also describes video applications and use of the method to multiply two signals; however, the original audio application is used for the purpose of illustrating the details of the operation of the method.

Various techniques have been used to control the output of electronic amplifiers in the face of variations of the amplitude of the input signal averaged over an arbitrary interval of time. This problem is particularly familiar to engineers in the fields of audio and of radio broadcasting, and more recently, in television broadcasting. Without enumerating each of the many uses to which the present method can be put, the authors merely wish to point out that in radio transmission, for instance, a standard level of signal strength is obviously desirable from the viewpoint of optimum utilization of the transmitting facilities; automatic control of the occasional excessive modulation peaks permits a substantial increase in the signal power. In audio reproduction, the new method enables the listener to restore the full contrast present in the original signal (e.g., in vocal and instrumental music), which is generally reduced by compression introduced at the origin.

Most of the circuits that have been proposed for controlling the amplification of a signal suffer from one or more serious defects. One such defect is that the control voltage may appear in the output of the amplifier and, moreover, may be many times larger than the desired signal itself. Some attempts have been made to reduce this undesirable effect by placing a limitation on the speed of response of the control operation so that the frequency of the control voltage lies well below the frequency of the signal voltage and may be thus filtered out relatively easily. Alternately, push-pull circuits or balanced-modulator circuits have been used to cancel out the control voltage in the amplifier output. However, such remedies may restrict the control circuits to relatively low speeds of response and require expensive transformers and dynamically balanced tubes; moreover, it is usually necessary to operate amplifiers of this type at very low volume levels to maintain distortion within tolerable limits. As a result, previously proposed methods have been slow in operation, restricted in bandwidth, delicate in balance, and subject to considerable distortion.

The present method suffers from none of these difficulties. The method consists of generating a selected carrier, modulating that carrier in accordance with the desired signal, modulating the signal-modulated carrier in accordance with a control voltage that may be derived from the desired signal itself or from an independent source, and demodulating the doubly modulated carrier simultaneously with respect to both modulations in such a manner that the control voltage appears only in combination with the desired signal but is not demodulated by itself. It is found that the effects of dynamic interaction between FM and AM signals are negligible.

Expansion and Compression

Expander

To obtain expansion operation, the incoming signal is used to frequency-modulate a local oscillator (Fig. 1).[3] The resultant FM voltage is amplitude-limited if neces-

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[1] W. R. Aiken, U. S. Patent No. 2,923,887; February 2, 1960.

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[3] Phase modulation may also be used, similarly to the combination of amplitude and phase modulation described, for instance, in D. G. Fink, ed., "Television Engineering Handbook," McGraw-Hill Book Co., Inc., New York, N. Y., Sec. 13.104; 1957.

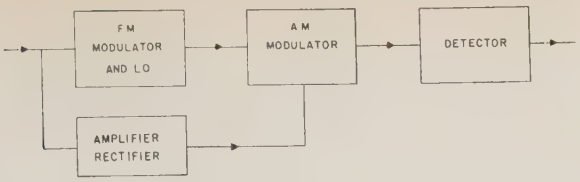

Fig. 1—Expander operation. Incoming signal frequency-modulates the local oscillator (typically a 5-Mc oscillator) and is simultaneously rectified in another circuit to provide the control signal. The FM signal is amplitude-modulated by the control signal to provide a composite signal, which is the FM signal with its amplitude *increased* by the control signal. The detector then yields the original signal with an increased relative-amplitude variation. No trace of the control signal remains.

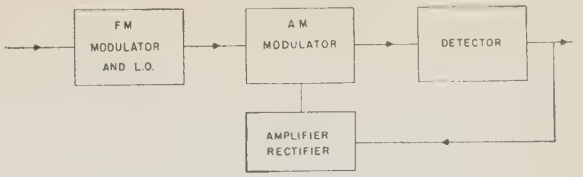

Fig. 2.—Compressor operation. Incoming signal frequency-modulates the local oscillator as in expander operation, but the control signal is now derived from the detector output and serves to decrease the amplitude of the FM signal to provide the composite signal. The detector then yields the original signal with the decreased relative amplitude variation. No trace of the control signal remains.

sary (to remove residual AM) and then amplitude-modulated by a control voltage that is also derived from the incoming signal; the control voltage is proportional to the (averaged) amplitude of the incoming signal. The output of the amplitude modulator is thus a signal that is both frequency- and amplitude-modulated; the FM content is proportional to the instantaneous value of the signal, and the AM content is proportional to the average value of the signal, as rectified in accordance with the delay bias. This bias is applied to the rectifier and serves to make its output nonlinear by shifting the diode cutoff point; the resulting increase in required signal level also changes the slope of the rectifier characteristic.

This composite signal is next detected by a Foster-Seeley or similar circuit responsive to both FM and AM, so that a demodulated signal results whose amplitude corresponds to the original signal as modified by the control signal. In other words, the output of the detector stage varies linearly with the instantaneous frequency deviation (FM) and the amplitude of the deviated carrier (AM). Since the detector cannot respond to AM in the absence of FM, only the modulating effect of the controlling signal (AM) is present in the output; the modulating signal itself does not appear. However, the restored signal now has a considerably greater dynamic range than the original signal.

Compressor

Operation as a compressor is illustrated in Fig. 2 and is substantially the same as the expansion action, except that the control signal is reversed in polarity and is derived from the output of the detector rather than from the input signal. Alternately, an independent source of control voltage may be used.

In audio applications, the method may be used in amplifiers in a relatively simple circuit for compression, limiting, or expanding. It is simpler, inherently more trouble-free, and has lower distortion and wider control range than existing circuits. Several versions of the compressor-expander for audio purposes have been demonstrated (Fig. 3), including a simplified, wholly transistorized one. Typical operating characteristics for action as expander and compressor are shown in Figs. 4 and 5 (see p. 1552), respectively. The effects of using the above mentioned delay bias are illustrated in both plots.

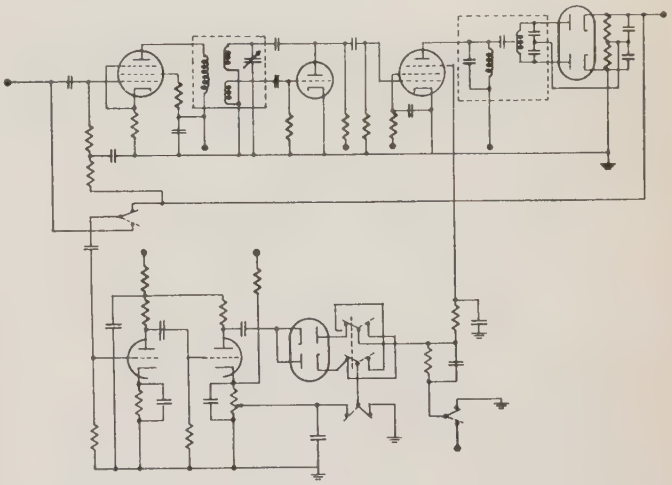

Fig. 3—Circuit diagram of early expander-compressor-limiter. Saturable reactor is used in the tuned circuit of the local oscillator (second tube) to effect frequency modulation. Later models are simplified by use of a varactor diode to replace the first (FM modulator) stage.

Video Applications

Although the method is being used to date at audio frequencies, it is inherently capable of operation at video frequencies as well. In video applications, the method can be used not only for controlling average signal levels, but for responding to individual picture elements. In this way, the shadow may be taken out of a video display or a scene may be brought out of partial shadow while another portion of the display is at high intensity. Moreover, the large dynamic range of vidicon camera tubes can be fully utilized without the danger of overloading their amplifier circuits. Such automatic control, in conjunction with voltage feedback to the vidicon tube, permits the use of high (and varying) light levels while background or average levels are maintained constant. This application can thus result in a constant-contrast display that would have numerous applications in commerical, industrial, and military telecasting.

Computer and Other Applications

The method is also applicable to a number of other uses, as in instrumentation, remote control, and indeed any uses involving modulation with separated components. It is capable of attenuating a signal by as much as 60 db without the distortion that occurs in conventional

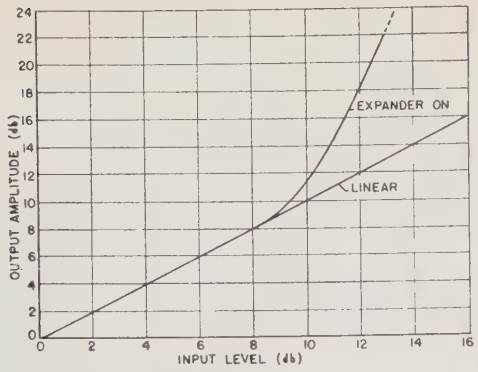

Fig. 4—Operating characteristics of expander: EXPANDER OFF yields linear characteristic, EXPANDER ON yields curved characteristic. Amount of control bias (in a back-biased diode) shifts the level at which expansion sets in (*i.e.*, the point at which the two curves diverge).

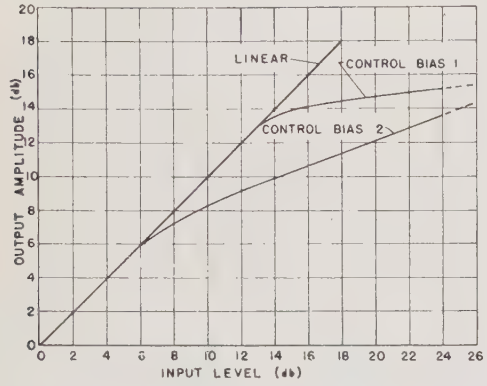

Fig. 5—Operating characteristics of compressor. Action opposite to that shown in Fig. 3 is obtained. (Two values of control bias are shown.)

circuits with nonlinear transfer characteristics. One use that holds particularly good promise of widespread practical application is as an electron *multiplier* in analog computers and similar devices. This action is demonstrated in Fig. 6.

It will be appreciated that the method described constitutes a linear electronic multiplier in which two voltages are multiplied together to obtain a product. The FM modulating voltage and the AM modulating voltage are the multiplying factors and the detector output is the product. The multiplicand may be used to frequency-modulate an RF oscillator, the multiplier may amplitude-modulate the FM signal, and the signal that is detected according to the described method represents the product.

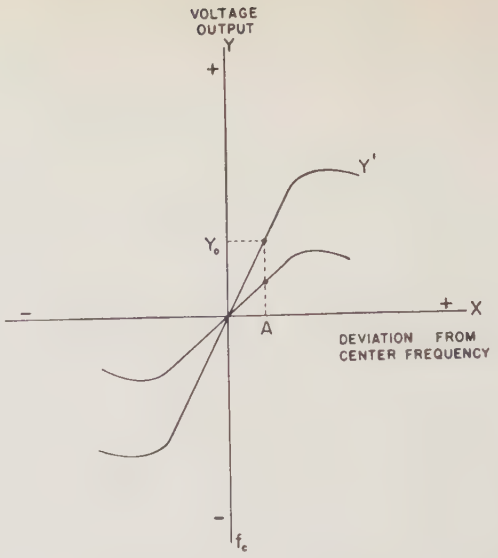

Fig. 6—Action of multiplier circuit. Multiplicand A is represented by deviation from center frequency, multiplier B, by the *slope* of the detector curve Y' (two values are shown).

The response curves of the detector are illustrated in Fig. 6. The amount of the multiplicand A, positive or negative, determines the position of the RF signal on the detector curve (positive or negative) as $+X$ or $-X$. The amount of the multiplier B determines the output-voltage amplitude, since the steepness of the detector curve depends on the amplitude of the RF signal.

If the multiplier $B=0$, the amplitude of the RF wave is zero, and the output will be along the line $X=0$. A nonzero value of the multiplier B results in an RF wave of corresponding amplitude, and the detector curve of Y' results, giving an output voltage of Y_0. In this way, the multiplicand A moves the RF signal along a detector curve whose slope is determined by the multiplier B. It should be noted that the output of the device is sensitive to both positive and negative values of the multiplicand A, and that the multiplicand and multiplier may be interchanged.

Conclusion

A new method of using the combination of two signals has been proposed and demonstrated by which compressor-expander action can be obtained without the disadvantages of existing schemes. The method is inherently capable of being used at video as well as audio frequencies, and is suitable for many other applications, notably in a multiplier circuit for computer applications

IRE Recommended Practices on Audio and Electroacoustics: Loudspeaker Measurements, 1961*

61 IRE 30.RP1

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1. Scope

These Recommended Practices define terms associated with loudspeakers and their testing, recommend various methods of testing, and indicate preferred methods of presenting information regarding their characteristics. Specific information is presented in Sections 2–7. Discussions of a more qualitative nature are given in Sections 8 and 9.

In these Practices, the tests recommended involve physical, steady-state measurements only. Work has been and is now being done on transient measurements of loudspeaker performance, but experience with these methods is still not sufficiently widespread to warrant their inclusion.

While the physical data which can be obtained as detailed in Sections 4–9, inclusive, are a helpful guide in designing and in selecting a loudspeaker for a certain purpose, they are not a complete guarantee that the subjective performance will be satisfactory. Wherever it is possible, the quality of reproduction should be checked by means of listening tests such as those described in the literature.[1]

2. Introduction

2.1 Definition

A loudspeaker is an electroacoustic transducer intended to radiate acoustic power into the air.

2.2 Types

There are two general types of loudspeakers, namely, direct-radiator and horn. The direct-radiator type of loudspeaker is arranged so that it radiates acoustic power into the air directly from one or both sides of its acoustic radiating element, or diaphragm. One side of the diaphragm may be totally isolated by means of an airtight enclosure, or, effectively, by a baffle of adequate size. In order to control the response over part of the useful frequency range, the energy radiated from the back surface of the diaphragm may be acoustically coupled to the energy radiated from the front.

The horn type of loudspeaker radiates the major portion of its acoustic power through a horn. This is a tube whose shape and cross section are chosen by design so as to increase the radiation resistance loading on the diaphragm of the loudspeaker for a substantial portion of the useful frequency range, and increase its radiated output and electroacoustic conversion-efficiency over that obtainable when it radiates directly.

Both direct-radiator and horn loudspeakers have been designed for a wide variety of applications which influence their range, directional characteristics, permissible distortion, and power handling capacity. Also, either type may be designed to operate as a single or multiunit device. One loudspeaker may be used to cover the desired frequency range. Alternatively, two or more loudspeakers, usually interconnected by means of frequency-selective electric networks, may be operated together so that each covers a portion of the desired frequency range, while the combination of loudspeakers covers the entire range.

3. Measurement Conditions

3.1 Loudspeaker Mounting

The direct-radiator loudspeaker unit should be mounted in or on the enclosure or baffle called for by the manufacturer, or it should be mounted in or on a baffle or enclosure which is completely described. The horn loudspeaker unit should be equipped with a horn specified by the manufacturer, or with a horn which is completely described.

3.2 Electrical Connections

The loudspeaker should be connected as specified by the manufacturer, with the proper networks and controls, if any, and these networks and controls should be adjusted as recommended. If not specified, the connections and adjustments used must be completely described.

If the loudspeaker is of the excited-field type, the field should be energized with the rated current.

3.3 Acoustic Environment

The acoustic environment should simulate free-field conditions to the extent that the inverse-pressure vs distance law should hold within plus or minus 1 db at all frequencies at which measurements are made.[2] Ambient noise should not affect the measurements to an extent greater than plus or minus 1 db. Departures from these conditions should be measured and described.

3.4 Test Signals

Fixed frequency, swept frequency and warbled sinusoidal signals, as well as bands of noise, may be used for loudspeaker testing as dictated by the test and the type of instrumentation being utilized. Sweep and warble rates should be sufficiently slow that response-frequency measurement results do not differ significantly from the steady-state readings. When bands of noise or warble signals are used their characteristics should be specified.

Each performance parameter of a loudspeaker must be expressed as a function of frequency if it is to be expressed completely. However, for many purposes it is desirable to condense the information to a single number which represents a weighted average over a prescribed portion of the spectrum. Such an average may be obtained by means of a signal which sweeps re-

[1] H. F. Olson, "Subjective loudspeaker testing," IRE Trans. on Audio, vol. AU-1, pp. 7–9; September–October, 1953.

[2] H. F. Olson, "Acoustical Engineering," D. Van Nostrand Co., Inc., New York, N. Y., 3rd ed., ch. 10, pp. 445–450; 1957.

peatedly through the required portion of the spectrum at a rate sufficient to give a steady reading on a suitable indicating device.[3]

The electric power input for pressure-frequency response, pressure-directional response, and impedance measurements, should be kept as low as possible consistent with the ambient noise and the sensitivity of the measuring system, in order to minimize the effects of heating and nonlinear distortion in the loudspeaker. However, when it is desired to test a loudspeaker at high levels, adequate time should be allowed for the attainment of steady-state thermal conditions. Likewise, when testing excited-field loudspeakers, steady-state thermal conditions should be attained.

3.5 Measuring System

The response of the measuring system including the microphone should be known. Nonlinear distortion should be limited to a level consistent with the type of data being obtained; however, moderate levels of system distortion are permissible for tests of maximum power capacity based upon mechanical or electrical failure. For nonlinear distortion measurements, the level of distortion products in the measuring system should have a negligible effect on the readings of distortion products produced by the loudspeaker. The reference-pressure response (fundamental) and hence, the pressure-frequency response (fundamental) by definition require measurement of the pressure at the fundamental frequency. A frequency-selective indicating or recording system may be required in order to minimize errors arising from harmonics in the output sound wave.

The microphone used for measuring the sound pressure should preferably conform to the standards given in American Standard Specification for Laboratory Standard Pressure Microphones, Z24.8-1949, or the latest revision thereof, and be calibrated by the method described therein. Other types of microphones may be used if calibrated as described in American Standard Method for the Free-Field Secondary Calibration of Microphones, Z24.11-1954, or the latest revision thereof. If the microphone employs a pressure-gradient element, it is necessary to correct for proximity effect at low frequencies.[4]

The measuring device used for single frequency measurements should provide readings which represent the rms values. For complex waves or bands of noise, the readings should be obtained by means which indicate peak and rms values for the type of complex wave used[5] or by a means with which these can be accurately correlated. When an automatic recording device is used, the rate of frequency sweep should be slow enough so that the trace does not differ significantly from the corresponding single frequency reading.

3.6 Preconditioning

Permanent changes may take place in a loudspeaker as a result of motion of the diaphragm. It is therefore desirable to subject it to a period of preconditioning by applying a signal of the type discussed in Section 9. The power of the signal should be the average value at which the loudspeaker will be used. The preconditioning should continue until changes of impedance, principal resonance and response can no longer be detected.

4. Electric Impedance Characteristics

Four electric impedances are of importance in testing loudspeakers. They concern the loudspeaker itself and the source from which it receives electric energy. It is important to determine their values in order that the loudspeaker may be tested under proper conditions with regard to such characteristics as power transfer, frequency response, and nonlinear distortion.

4.1 Electric Impedance of a Loudspeaker ($\boldsymbol{Z}_S$)

4.1.1 Definition. The electric impedance of a loudspeaker is the complex value of the electric impedance given as a function of frequency and measured at the accessible signal terminals of the loudspeaker.

4.1.2 Method of Measurement. The loudspeaker should be mounted, connected and tested in a suitable acoustic environment as discussed in Section 3.3. The impedance should be determined from measurement of magnitudes and phase relationship of the voltage and current supplied to the loudspeaker terminals.

Note: The impedance measurement should be made at an operating level at which voltage and current waveforms are essentially sinusoidal.

4.1.3 Presentation of Data. Loudspeaker impedance data should be presented in complex form, indicating magnitude and phase angle as a function of frequency, or alternatively, resistive and reactive components as a function of frequency. Electric input conditions should be specified.

4.2 Rating-Impedance of a Loudspeaker (Z_R)

4.2.1 Definition. The rating-impedance of a loudspeaker is the value of a pure resistance which is to be substituted for the loudspeaker when the electric power supplied by the source is to be measured.

Note 1: This is sometimes referred to as nominal impedance.

Note 2: As an example, a loudspeaker having a designated rating-impedance of eight ohms is intended to be connected to the eight-ohm output tap of an amplifier.

[3] H. F. Hopkins and N. R. Stryker, "A proposed loudness-efficiency rating for loudspeakers and the determination of system power requirements for enclosures," Proc. IRE, vol. 36, pp. 315–355; March, 1948.

[4] Olson, *op. cit.*, ch. 8, p. 288 and Fig. 8.40A.

[5] L. L. Beranek, "Acoustic Measurements," John Wiley and Sons, Inc., New York, N. Y., ch. 11; 1949.

4.2.2 Evaluation. It should be noted that the rating-impedance is not the impedance of the loudspeaker at a designated reference frequency. It is, ideally, the average impedance over the frequency band transmitted by the loudspeaker, weighted by the spectrum of the signal with which the loudspeaker will be used, and by such factors as distortion and frequency response.

Note: The manufacturer usually recommends a value of rating-impedance, but when this information is lacking it may be estimated by one of the following procedures:[6]

a. For moving-coil direct-radiator loudspeakers, use the minimum value of the magnitude of the measured loudspeaker impedance in the frequency range above cone resonance, and add 10 per cent; or, measure the voice-coil dc resistance and add 20 per cent.

b. For horn loudspeakers in general, use the average of the magnitude of the measured loudspeaker impedance in the region at the center of the useful frequency band of the unit; or, as an alternative in the case of moving-coil horn loudspeakers, measure the voice-coil dc resistance and add 40 per cent.

4.3 Loudspeaker Measurement Source Impedance (Z_G)

4.3.1 Definition. The loudspeaker measurement source impedance is the value of a pure resistance to be connected in series with the loudspeaker and a constant-voltage source in order to measure the loudspeaker performance.

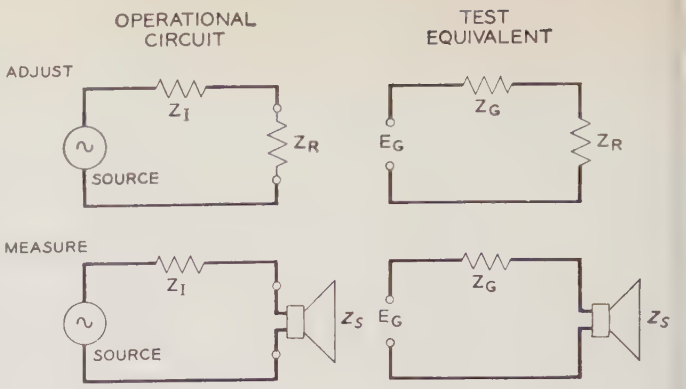
Note: The choice of the value of Z_G will depend upon the type of amplifier to which the loudspeaker will be connected in use.

4.4 Discussion of Impedances

The efficiency of a loudspeaker varies as a function of frequency, and depends not only upon its ability to transform electric to acoustic power, but also on its ability to accept power from a source. The source may be a power amplifier, from which maximum power consistent with prescribed distortion conditions may be obtained only with a load impedance that is not necessarily equal to the source impedance. In order to reduce the discrepancies between idealized measurements with resistive sources and actual use with an amplifier, it is necessary to specify the four impedances involved in these two processes: the three defined in this section $\boldsymbol{Z}_S$, Z_R and Z_G, and a fourth, $\boldsymbol{Z}_I$, the internal impedance at the output of the amplifier with which the loudspeaker is to be used.

The use of these quantities is illustrated in Fig. 1.

The relation between generator and load impedance may be expressed by the regulation of the combination,



E_A is the rms value, in volts, of the open-circuit voltage of the amplifier.
E_G is the rms value, in volts, of the open-circuit voltage of the source.
Z_G is the loudspeaker measurement source impedance, in ohms.
$\boldsymbol{Z}_I$ is the internal impedance, in ohms, at the output of the amplifier with which the loudspeaker is to be used.
Z_R is the loudspeaker rating-impedance, in ohms.
Z_S is the impedance, in ohms, of the loudspeaker, measured at its terminals.

In the "ADJUST" row, the electric power P_E delivered to the loudspeaker rating-impedance, Z_R, is set by adjusting the source voltage E_G. This electric power delivered to Z_R is taken as a measure of the input power available to the loudspeaker.

Fig. 1.

where

$$\text{Regulation in db} = -20 \log_{10} \frac{\text{(terminated-circuit output voltage)}}{\text{(open-circuit output voltage)}}.$$

It is important that the regulation be stated along with other conditions of measurement.

Note: For example, a properly terminated triode amplifier has a regulation of approximately 3 db; for a matched source-load combination the corresponding regulation becomes 6 db; for a constant voltage generator, 0 db.

5. Reference Pressure Response Characteristics

5.1 Reference Pressure Response of a Loudspeaker (G_L)

5.1.1 Definition. The reference pressure response of a loudspeaker at a specified frequency, expressed in decibels, is the normalized pressure response indicating the ratio of output sound pressure level at that frequency in a free field on the principal axis referred to a distance of one meter, to the input electric power level at that frequency delivered to a resistance equal to the loudspeaker rating-impedance.

Units: The recommended reference quantities are 0.0002 microbar (dyne per cm^2) for pressure and 0.001 watt for power. For convenience, the reference pressure

[6] V. Salmon, "Loudspeaker impedance," IRE Trans. on Audio, vol. AU-1, pp. 1–3; July–August, 1953. See also "Coupling the speaker to the output stage," Newsletter of the IRE-PGA, no. 3, p. 5; January, 1952.

response G_L, in db, may be expressed in the following forms:

$$G_L = h - H = 20 \log_{10} (p_L/p_0) - 10 \log_{10} (P_E/P_0)$$
$$= 20 \log_{10} p_L - 20 \log_{10} E_G + 10 \log_{10} Z_R$$
$$+ 20 \log_{10} (1 + Z_G/Z_R)$$
$$- 20 \log_{10} p_0 + 10 \log_{10} P_0$$

and, on substitution of recommended reference quantities, the expression

$$-20 \log_{10} p_0 + 10 \log_{10} P_0 = 44 \text{ db},$$

where

h is the axial free-field sound pressure level produced by the loudspeaker at a distance of 1 meter, in db referred to 0.0002 microbar (dyne per cm^2).
H is the electric power level delivered to the rating-impedance, in dbm (db referred to 0.001 watt).
p_L is the axial free-field rms sound pressure at a distance of one meter, in microbars.
p_0 is the reference sound pressure, 0.0002 microbar.
P_E is the electric power, in watts, delivered to a resistance equal to the rating-impedance of the loudspeaker, where

$$P_E = \frac{E_G^2 Z_R}{(Z_G + Z_R)^2}.$$

P_0 is the reference electric power, 0.001 watt.
E_G is the rms value, in volts, of the open-circuit voltage of the source.
Z_R is the loudspeaker rating-impedance, in ohms.
Z_G is the loudspeaker measurement source impedance, in ohms.

5.1.2 Method of Measurement. The loudspeaker should be mounted, connected, and tested in a suitable acoustic environment as discussed in Section 3.

The microphone should be placed on the principal axis of the loudspeaker as specified by the manufacturer, or, if no axis is specified, on the normal to the center of the radiating area or horn mouth, and at a distance at least three times the maximum transverse dimension of the radiating area or horn mouth in order to approximate free-field conditions.

From the free-field pressure p_r, measured at a distance D(feet), the pressure p_L at one meter can be computed by the relation:

$$p_L = p_r(D/C_1),$$

or, in logarithmic form

$$p_L \text{ in db} = 20 \log_{10} p_r + 20 \log_{10} D - 20 \log_{10} C_1,$$

where

$$C_1 = 3.28 \text{ (number of feet in one meter)}.$$

5.2 Average Reference Pressure Response

When it is desired to rate a loudspeaker by means of a single value of the reference pressure response, it is preferable to take an average over a frequency band[3] rather than to cite the value obtained for any one frequency. The method used should be described.

5.3 Reference Pressure-Frequency Response

5.3.1 Definition. The reference pressure-frequency response of a loudspeaker is the reference pressure response presented as a function of frequency.

Note: The relative pressure-frequency response characteristic, commonly known as frequency response, is obtained by referring the response values at various frequencies to that at some reference frequency.

5.3.2 Presentation of Data. When presenting data on the reference pressure-frequency response of a loudspeaker, the following quantities should be specified: measured distance between the loudspeaker radiating area or horn mouth and the microphone; electric power input to the loudspeaker rating-impedance; value of this loudspeaker-rating impedance; and value of the measurement source impedance.

Note: Unless there is a special reason for using other types of graphic presentation for pressure-frequency response curves, it is recommended that the frequency scale, or abscissa, be logarithmic, and the pressure scale, or ordinate, be linear in db, and that the length of a 10 to 1 frequency interval be the length of 30 db on the ordinate scale.

6. Rating-Efficiency (η)

6.1 Definition

Rating-efficiency is the ratio, usually expressed in decibels, of the output acoustic power to the electric power delivered to a resistance equal to the loudspeaker rating-impedance, that is,

$$\eta = 10 \log_{10} (P_L/P_E),$$

where

η is the loudspeaker rating-efficiency, in db.
P_L is the total radiated acoustic power, in watts.
P_E is the electric power, in watts, delivered to a resistance equal to the rating-impedance of the loudspeaker, where

$$P_E = \frac{E_G^2 Z_R}{(Z_G + Z_R)^2}.$$

6.2 Discussion

The acoustic output of a loudspeaker can be determined under free-field conditions by combining a measurement of the pressure on the reference (usally principal) axis with the directivity index. The formula for

loudspeaker rating-efficiency η, in db, may be expressed in these terms as follows:

$$\eta = 20 \log_{10} p_L - 20 \log_{10} E_G + 10 \log_{10} Z_R + 20 \log_{10} (1 + Z_G/Z_R) - K_L - C$$

where

K_L is the loudspeaker directivity index, in db (see Section 7.1.3).

C is a constant which relates the pressure, in microbars at a distance of one meter to the acoustic power, in watts, transmitted through a sphere of 1 meter radius. For air, under standard conditions of temperature and barometric pressure, the value of this constant C is 35.2.

Note: Another method of determining loudspeaker efficiency which does not require a measurement or estimate of directivity index, utilizes a reverberant room.[7]

7. Directional Properties

In many cases, the directional characteristics of a loudspeaker can be sufficiently described by measuring one or more pressure-frequency responses along axes at chosen angles to the normal axis as specified by the manufacturer. These will indicate how well the loudspeaker response characteristics obtained on the principal axis are duplicated over an area in which listeners may be located for a given application.

In order to determine the loudspeaker rating efficiency when operating under free-field conditions, it is necessary to compute the directivity index. This requires that pressure-frequency responses be obtained at a large number of angles, or pressure-angle responses at a large number of frequencies, to provide adequate data. For many applications it is sufficiently accurate to use computed instead of measured values of the directivity index.[8] An average value, using a test signal which covers a wide band of frequencies, has been found useful in public address and sound systems engineering.[3]

7.1 Directivity Factor and Directivity Index of a Transducer

7.1.1 Directivity Factor—Definition. The directivity factor of a transducer used for sound emission is the ratio of the intensity of the radiated sound at a remote point in a free field on the principal axis, to the average intensity of the sound transmitted through a sphere passing through the remote point and concentric with the transducer. The frequency must be stated.

Note: The point of observation must be sufficiently remote from the transducer for spherical divergence to exist. See Fig. 2.

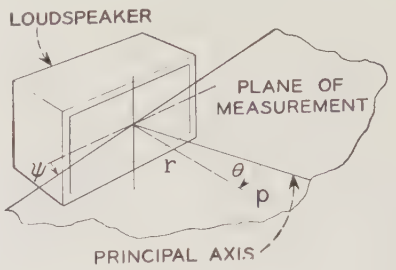


Fig. 2.

7.1.2 Directivity Index—Definition. The directivity index of a transducer is an expression of the directivity factor, in decibels, *viz.*, ten times the logarithm to the base ten of the directivity factor.

7.1.3 Directivity Index—Mathematical Expression. The directivity index may be expressed in the following form:

$$K_L = -10 \log_{10} \frac{1}{4\pi} \int_0^{2\pi} \int_0^{\pi} (p/p_r)^2 \sin \theta d\theta d\psi,$$

where

K_L is the loudspeaker directivity index, in db.

p_r is the axial free-field sound pressure, in microbars, at distance r.

p is the free-field sound pressure, in microbars at the point (r, θ, ψ).

θ and ψ are the angular polar coordinates of the system, and the principal loudspeaker axis is at $\theta = 0$.

7.1.4 Directivity Index—Method of Measurement. The loudspeaker should be mounted, connected, and tested in a suitable acoustic environment as discussed in Section 3.

The reference axis for the position of the microphone should be the principal axis of the loudspeaker as specified by the manufacturer, or, if no principal axis is specified, the normal from the center of the radiating area or horn mouth. The microphone should be placed at a distance at least three times the maximum transverse dimension of the radiating area or horn mouth. The angular steps should be as small as needed to delineate accurately the directional pattern in the plane chosen. If the loudspeaker has no axis of symmetry, the measurements should be repeated in other planes until the field is mapped.

The directivity index may then be calculated by numerical integration of the expression given under the definition above.[9]

[7] H. C. Hardy, H. H. Hall, and L. G. Ramer, "Direct measurement of the efficiency of loudspeakers by use of a reverberation room." IRE Trans. on Audio, No. AU-10, pp. 14–24; November-December, 1952.

[8] C. T. Molloy, "Calculation of the directivity index for various types of radiators," *J. Acoust. Soc. Am.*, vol. 20, pp. 387–405; July, 1948.

[9] "Calibration of Electroacoustic Transducers," ASA Z24.24; 1957. Contains information helpful in the calculation of directivity index.

8. Nonlinear Distortion

8.1 Nonlinear Distortion in a Loudspeaker

8.1.1 Definition. Nonlinear distortion in a loudspeaker is an undesired change in waveform due to deviation from a linear relationship between acoustic output and electric input.

8.1.2 Discussion. The effect of this distortion on quality constitutes one of the factors which limit the useful magnitude of the output of a loudspeaker. Some of the more important causes of this type of distortion are: (1) nonlinear force-displacement relationships in the mechanical vibrating system of the loudspeaker, (2) nonuniform flux linkage throughout the displacement range in moving-coil loudspeakers, and (3) nonlinearity in the air itself, especially in horn loudspeakers. The first two causes produce significant distortion at low frequencies for which the displacement is greatest; the third cause is significant at high frequencies.

In consequence of nonlinearity, components having frequencies other than those appearing in the electric input signal will be found in the acoustic output. For a single frequency input, these nonlinear distortion products will be harmonics and subharmonics of the original input. In the case of a complex wave input there will be, in addition, intermodulation products, *i.e.*, sum and difference frequencies of the original components and of various multiples of these components.

Nonlinear distortion is commonly determined by a measurement of the distortion products, using a microphone to convert the acoustic output of the loudspeaker to a corresponding electric signal and appropriate networks to identify and measure its various components. Distortion products fall in various parts of the spectrum, having frequencies either above or below the fundamental signal or signals by a small or large frequency interval, and may be harmonically or nonharmonically related to the original signals that produce them. Each of these products may have a different subjective importance. Furthermore, loudspeaker distortion will vary markedly with the signal frequency or frequencies even with constant input power. For these reasons, the subjective evaluation of distortion through listening tests may be more useful than measured data; and for the same reasons, no single method of measurement of distortion is standardized herein.

8.1.3 Methods of Measurement. The loudspeaker should be mounted, connected and tested in a suitable acoustic environment as discussed in Section 3.

The microphone should be placed as specified in Section 5.1.2.

Each component of the output signal may be measured individually by means of a wave analyzer which provides a narrow pass band movable over the spectrum. The pressure amplitude of each component, expressed in per cent of the amplitude of the fundamental, may be plotted vertically above a frequency scale.

Harmonics of a single fundamental may be measured by the use of a high-pass filter which effectively suppresses the fundamental and passes the harmonics without attenuation. The result, as indicated by a square-law meter (see Section 3.5), will be the root-sum-square of all harmonics present. Subharmonics will not be included.

Subharmonics may be measured with the use of a low-pass filter which suppresses the fundamental and transmits all lower frequencies without attenuation. When measured with a square-law meter the result is the rms value of all subharmonics.

Two input signals of equal amplitude are recommended for measurement of intermodulation distortion.

A filter may be considered to "suppress effectively" a fundamental if it attenuates the fundamental to an amplitude at least 10 db below that of the distortion product closest to it in frequency.

In selecting fundamental frequencies for a measurement, it should be recognized that (a) the greatest measured distortion occurs most often at the lowest frequencies; (b) distortion due to nonlinearity of the air occurs to a significant degree only in loudspeakers coupled to horns and is greatest at the higher frequencies; (c) harmonics and sum-products of plural test frequencies may not be radiated if the fundamental frequency or frequencies approach the upper limit of the effective radiating frequency range of the system, and nonlinearity may therefore be important principally in the production of difference-products; (d) difference-products may fall below the cutoff of a loudspeaker or horn and thus escape measurement; (e) harmonics and sum-and-difference-products falling in the middle portion of the frequency range of the loudspeaker, or of the ear, may be much more important subjectively than those at the extremities of the range; (f) subharmonics may be more important subjectively than harmonics; (g) nonharmonically related distortion products may be more important subjectively than harmonically related distortion products.

8.1.4 Presentation of Data. Loudspeaker distortion must be expressed as a function of frequency if it is to be expressed completely. The electric input connections to the loudspeaker and the type of equipment used for measurement must be specified.

9. Rated Power (Handling) Capacity

9.1 Power Capacity Rating of a Loudspeaker

Because of the many factors which influence the power capacity rating of a loudspeaker, no general agreement has been reached on what physical measurements should be made nor on a method of weighting the numerical value so obtained to arrive at a single rated power capacity. There is, however, general agreement on the object of such a rating.

The objective is to obtain a power capacity rating which indicates to the user that the loudspeaker will

operate satisfactorily with an amplifier having the same power rating. Some of the more important factors to consider in the rating and in deciding whether operation is "satisfactory" are:

1) operating life,
2) freedom from spurious noises (buzzes, rattles, etc.),
3) acceptable nonlinear distortion for the intended application.

The operating life may be limited by electrical failure or mechanical failure arising from mechanical or thermal causes. The acceptable noise and nonlinear distortion depend on the application. For example, more noise and distortion might be tolerated in high-level, narrow-frequency-band, public-address speech systems than in broad-band, music systems. The standardization of power capacity ratings is further complicated by the fact that the important rating factors vary greatly with frequency and the frequency dependence differs in various designs intended for the same application. At present, subjective tests are an essential part of loudspeaker rating procedures. In making physical tests for rating procedures, the following minimum requirements should be considered.

9.1.1 Testing Arrangements. The loudspeaker should be mounted and connected to the test equipment as discussed in Sections 2.1 and 3.2. The acoustic environment should simulate approximately that in which the loudspeaker will be used. The test enclosure should be of sufficient size and should employ sufficient damping that no abnormal resonances occur which might lead to abnormal mechanical fatigue in the loudspeaker.

9.1.2 Type of Test Signal. The test signal should simulate the energy distribution, both in frequency and time, of the intended signal source material. This is important in order to excite all modes of vibration that may give rise to physical fatigue or noise and to insure normal and not excessive heating. This is particularly important in moving-coil loudspeakers where the rating may be limited by the temperature of the voice coil.

The rated power capacity of the loudspeaker is normally based on the type of signal obtained from a linear system. Short-time-peak to long-time-average power ratios may be in the order of 20 db in linear speech and music systems. If the system designer intends to reproduce a modified signal, for example, a speech signal altered by substantial compression and limiting or clipping in which both the frequency and time distribution of the energy are modified, a special rating will, in general, be required.

Because of the difficulty of producing a satisfactory artificial test signal, speech and music are frequently used. This is preferably selected recorded material.

Other test signals, such as pulsed sine wave, band of noise, warbled or swept frequency, have been used. The swept frequency signal is often one in which the frequency is varied linearly with time over a specified range.

Note: For accelerated life tests the signal is frequently supplied to the loudspeaker through a limiter so operated as to give approximately a 1/8 second peak to 15 second average power ratio of 6 db. This alters the energy distribution and substantially increases the heating effects. The rated loudspeaker capacity is said to be equal to the rated amplifier output power if no failure or significant performance degradation occurs in 100 hours. What is obtained relates to a system rating since the overload characteristic of the amplifier is also involved.

9.1.3 Duration of Test. The test should be continued for a sufficient length of time to permit (a) maximum steady-state temperatures to be developed at all points in the loudspeaker structure under the specified conditions of ambient air temperature and ventilation, and (b) fatigue failures of materials to develop (in order of 10^7 stress alternations).

Coincidence Techniques for Radar Receivers Employing a Double-Threshold Method of Detection*

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Summary—**The double-threshold method of detection calls for m or more detections in n consecutive trials to decide that a target is present. With the video coincidence techniques described, the requirement is m or more consecutive pairs, or triples, in n consecutive trials. The latter methods may increase the resistance to random interference, but a certain price has to be paid in normal radar-detection range. Important questions are, 1) If the interference has a probability p of exceeding the first threshold, what is the probability that a false target will be formed using these different techniques? 2) What is the required signal-to-noise ratio for a given probability of detection when the rate of false alarm is held constant for all three techniques? The paper develops mathematical formulas to answer such questions, and also presents graphs for the cases $n=10$, 20 and 30 hits per target. Examples are worked out for $n=10$.**

Introduction

THE performance of a radar set may be degraded by unintentional interference pulses from, *e.g.*, adjacent radar sets, or by intentional interference from a jamming source. The type of interference which will be considered in the following discussion is the one where random pulses appear at the detector output. To reduce the effect of such interference, advantage may be taken of the fact that pulses, due to a real-target return, appear at a constant position during several sweep periods, whereas pulses which are not locked to the radar-pulse repetition rate may occur at random. Therefore, as is well known, the effective interference may be reduced by correlating two or more sweep periods. What is not generally known is the quantitative advantage to be expected, and how such schemes will affect the normal radar performance.

The authors have tried to establish the necessary formulas for a radar set employing certain digital-detection techniques. It is assumed that the coverage area is quantized in azimuth and range, such that each quantum, or unit element, is identical in size and shape to one normal radar echo.

Basic Concepts

Single-Pulse Detection Probability

When a target is illuminated by a pulsed-radar set, the receiver will accept a number, say n, of returns during one scan period. Each pulse is more or less contaminated by noise, and the pulse envelope will therefore fluctuate. As an example, Fig. 1 shows the probability that the signal-plus-noise amplitude exceeds a given relative threshold R_t for various signal-to-noise ratios.[1] It is here assumed that the target echo itself is steady.

The target echo will in general also fluctuate. For a monofrequency radar system, the echo from a jet aircraft may, *e.g.*, be steady during one scan, but fluctuate from scan to scan. With a frequency-jumping radar set, an aircraft echo may also fluctuate considerably from sweep to sweep. Fig. 2 shows the probability of a single noise-contaminated radar return exceeding a given threshold R_t for the extreme case when the signal itself is also noiselike. Figs. 1 and 2 apply for a single pulse. In radar detection, however, an ensemble of n pulses is considered.

In PPI systems where an operator judges whether targets are present or not, the detection process is somewhat obscure, as the role of the operator as a detection element is not precisely understood. In mathematical analyses it is assumed generally that for every unit element each noise or signal-plus-noise sample is individually detected and all n detected samples added.[2,3] A target is judged to be present if the sum exceeds a certain threshold.

To account for the operator, an *operator degradation factor* is introduced, defined as the ratio between the experimental and theoretical signal-to-noise ratios required for detection. For example, this factor may be from two to ten times (3 to 10 db).[3]

Double-Threshold Method of Detection

For automatic data extraction, digital methods are more convenient, and the detection criteria can be more exactly defined. A usual model is shown in Fig. 3(a). The individual pulse is compared with a fixed threshold R_t (*first threshold*), and is classified as a one or a zero depending upon whether or not it exceeds the threshold. For each group of n pulses the number of ones are counted, and if the sum equals or exceeds a predetermined value m (*second threshold*), a target is judged to be present.

* Received by the IRE, May 5, 1961; revised manuscript received, June 2, 1961.

† Norwegian Defence Res. Establishment, Lillestrom, Norway.

[1] S. O. Rice, "Mathematical analysis of random noise," in "Selected Papers on Noise and Stochastic Processes," Dover Publications, Inc., New York, N. Y., pp. 238–241; 1954.

[2] E. L. Kaplan, "Signal-detection studies, with applications," *Bell Sys. Tech. J.*, vol. 34, pp. 403–437; March, 1955.

[3] W. M. Hall, "Prediction of pulse radar performance," PROC. IRE, vol. 44, pp. 224–231; February, 1956.

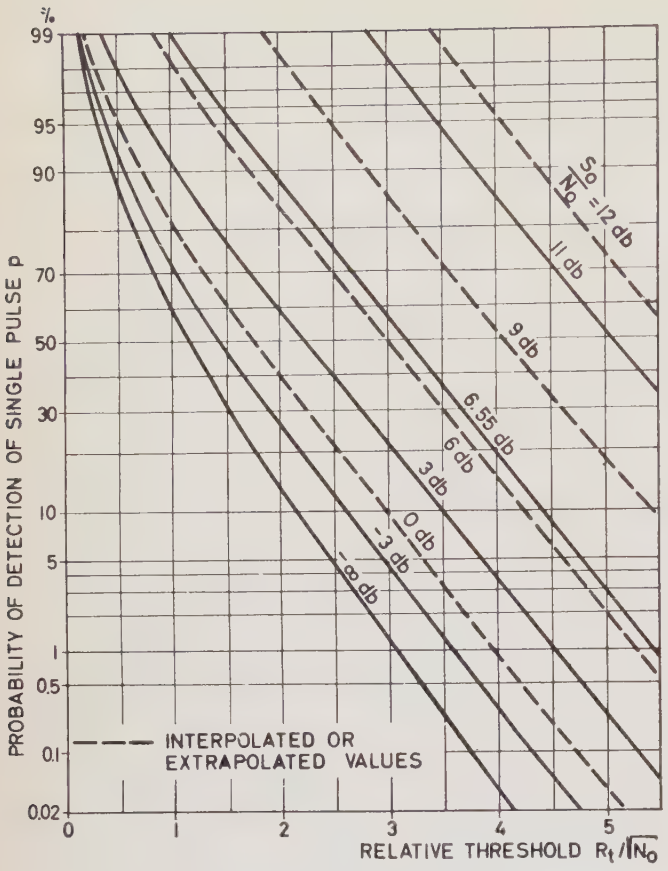

Fig. 1—Probability of detection of single pulse. No fading. S_0 = signal power. N_0 = mean noise power.

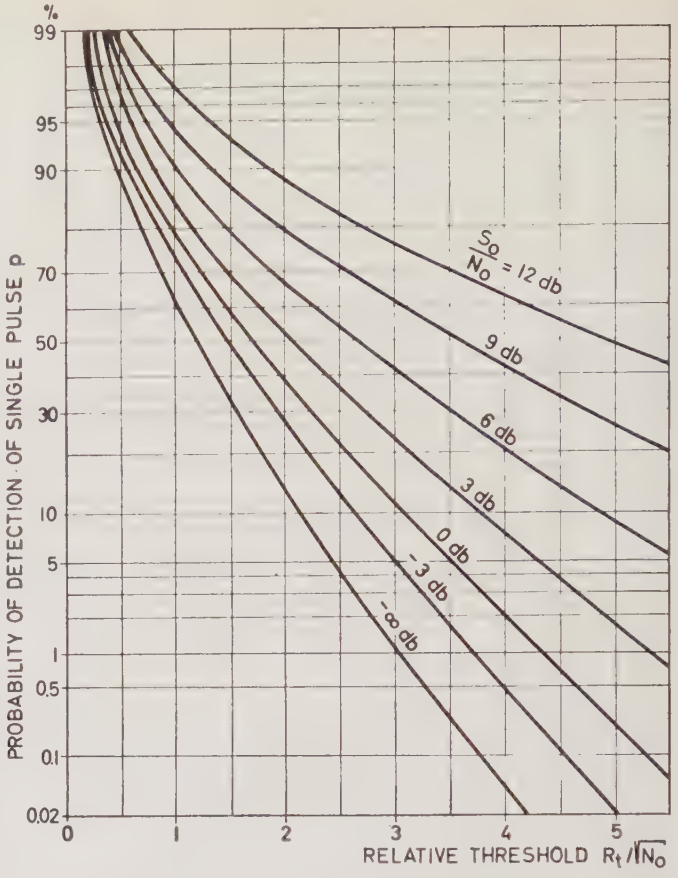

Fig. 2—Probability of detection of single pulse. Noise-like signal. S_0 = mean signal power. N_0 = mean noise power.

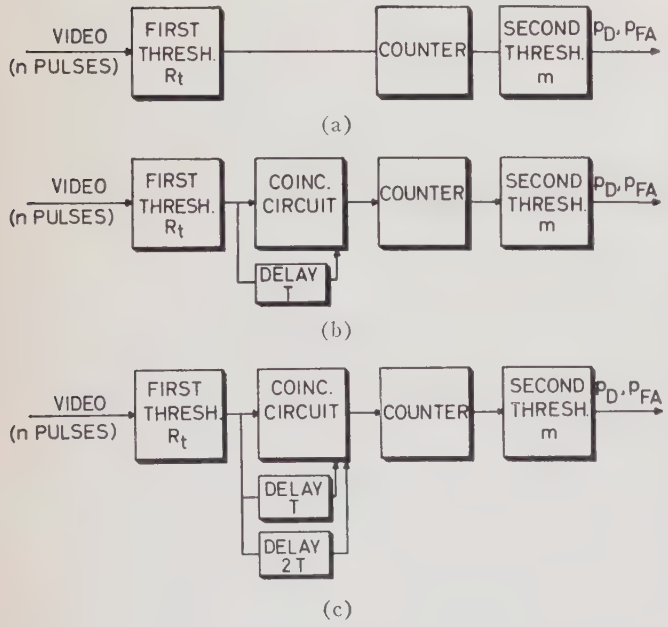

Fig. 3—Principle of coincidence circuits. (a) No coincidence. (b) Double coincidence. (c) Triple coincidence.

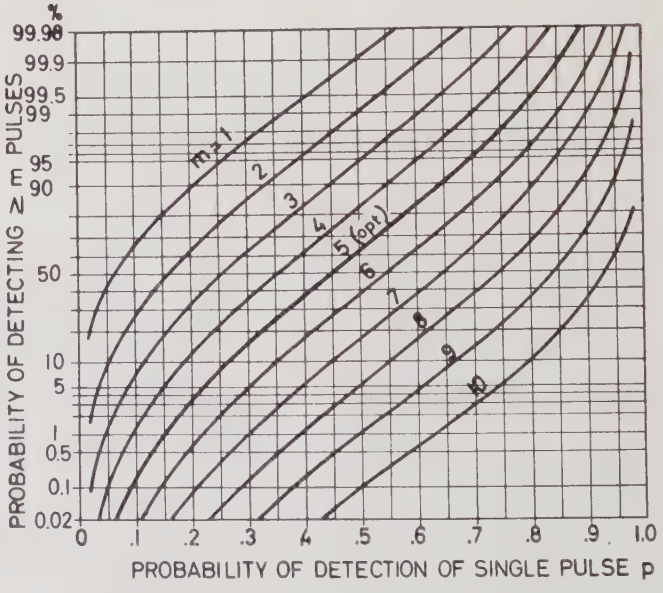

Fig. 4—Probability of detection in a system not employing coincidence techniques. $n = 10$.

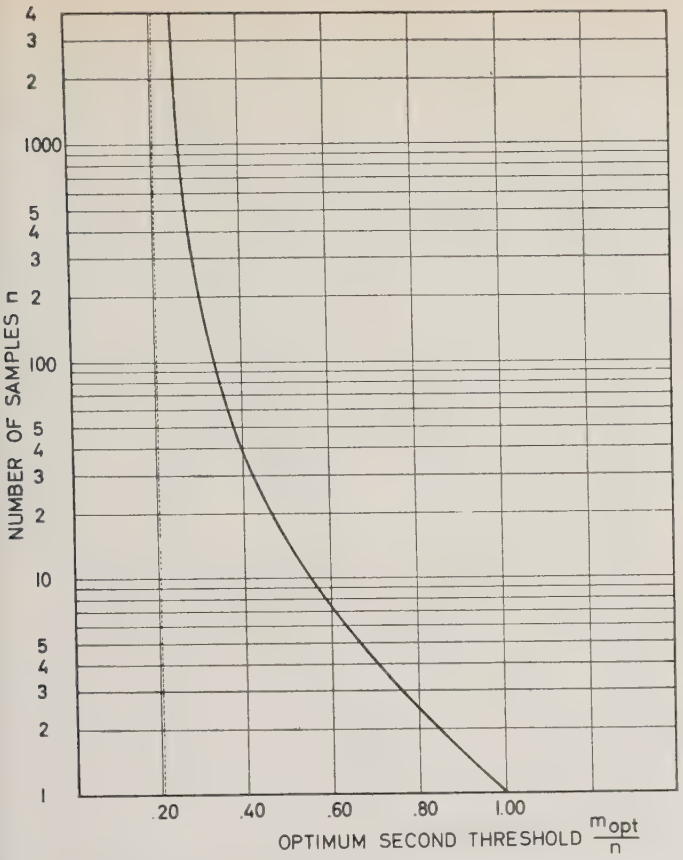

Fig. 5—Relation between the optimum second threshold and the number of samples.

Fig. 4 illustrates for the case $n = 10$ the dependence of detection probability upon the value of m and the single-pulse detection probability p. The latter value for a given signal-to-noise ratio may be obtained from Fig. 1 or Fig. 2.

Swerling[4] has shown that when m is suitably chosen, the double-threshold method requires about 1 db better signal-to-noise ratio than the post-detector integration method. An automatic system will be therefore in reality some 2–10 db better than a manual PPI system.

Fig. 5 gives (after Swerling) the approximate optimum value of the second-threshold m as a function of n.

False Alarm Rate

Whatever system is used, there is a possibility that random noise may be falsely judged as being due to a real target return. On a normal PPI oscilloscope the number of unit elements that can accommodate a target echo is of the order of 10^4–10^6. Consequently, if one requires that the mean number of false alarms occurring during one complete scan shall not exceed say one, the probability of false alarm within a given unit element should be less than, *e.g.*, 10^{-5}. This determines the thresholds to be used. In the double-threshold model, the optimum value of m is not very much influenced by the false alarm rate. The first threshold is more critical.

[4] P. Swerling, "The 'Double Threshold' Method of Detection," Rand Corp. Res. Memo. RM-1008; December, 1952.

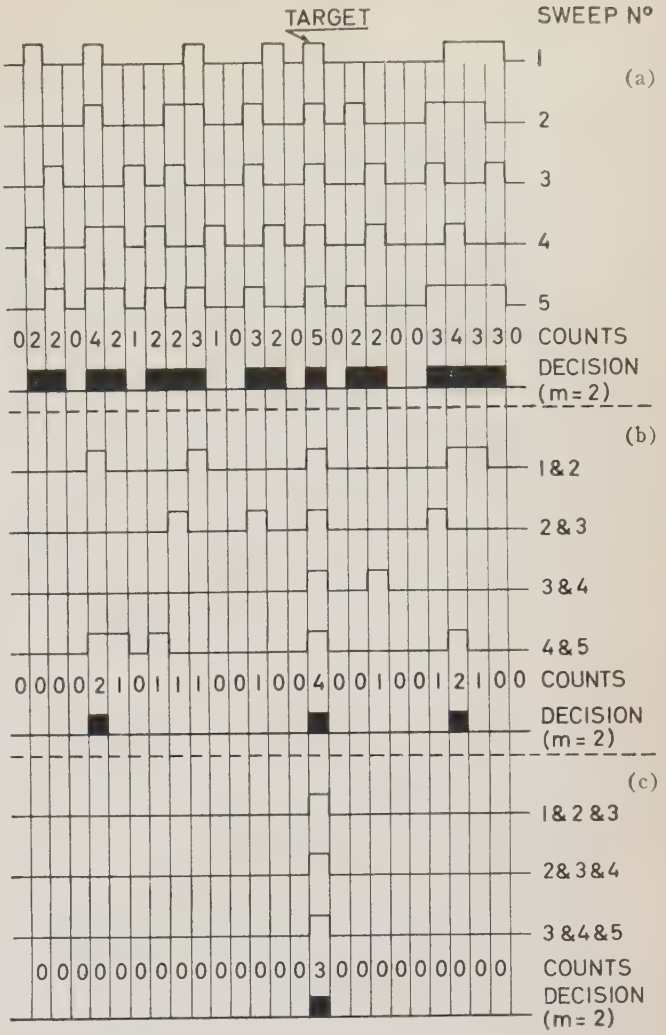

Fig. 6—Waveforms, example. (a) No coincidence (b) Double coincidence (c) Triple coincidence.

Video Coincidence Techniques

Description of Video Coincidence Circuits in a Double-Threshold System

To illustrate the principle of coincidence techniques in a double-threshold system, reference is made to Figs. 3 and 6.

In Fig. 3(a) (no coincidence) the video signal is first quantized as outlined above, and then passed to a number of counters (only one shown). Each counter only accepts pulses arriving within a certain range and azimuth interval. The range interval is equal to the duration of one normal radar pulse, and the azimuth interval equals the equivalent lobe width of the antenna. How this gating is performed will not concern the present discussion.

In Fig. 6(a) it is assumed that the number of sweeps per antenna lobe width is 5. For the purpose of illustration m is put equal to 2. One target is present; all other pulses are due to noise. The number of counts for each range interval and the corresponding decisions are indicated.

In Fig. 3(b) each sweep is coincided with the previous one. The corresponding waveforms are indicated in Fig. 6(b). The counts and decisions are indicated, assuming $m=2$ as before. In Figs. 3(c) and 6(c) each sweep is coincided with the previous two.

In the examples cited in Fig. 6, it appears that the number of false decisions is reduced as the number of coincidence inputs is increased. This is true if the interference is due to, *e.g.*, strong random-pulse jamming.

If the interference is due to noise, Fig. 6 is not representative. On the contrary, the first threshold would be set such that the number of false decisions was equal in all three cases, corresponding to the permissible false alarm rate. What would in fact vary in Fig. 6 is the detection probability for the real target.

Probabilities of Detection and False Alarm

The probabilities of detection and false alarm are derived in Appendixes I–III for the cases of one, two, and three coincidence inputs.

In the equations, the following symbols are used:

p_D = Probability of detection,
p = Probability that the received signal-plus-noise envelope exceeds the first threshold (*e.g.*, Figs. 1 and 2),
p_0 = Probability that the noise envelope exceeds the first threshold ($S_0/N_0 = -\infty$ db in Figs. 1 and 2),
p_{FA} = Probability of false alarm,
n = Number of pulses per target,
m = Second threshold. Approximate optimum values may be found from Fig. 5 (corresponding to n, $n-1$ and $n-2$ in the three cases).

No general conclusion may be drawn directly from the equations as they stand, they being solved better numerically for each case of interest. The NDRE electronic-computer FREDERIC has been programmed to give numerical values of p_D for any given n, m and p, employing the exact formulas (2), (5), and (7) presented in Appendixes I–III. Some results are given in Figs. 4 and 7–11. The optimum values of m are found from Fig. 5.

To indicate how the curves may be used, a few examples are given for the case $n=10$.

Example 1. Random Pulse Interference ($n=10$)

When the interference is due to nonsynchronized strong pulsed or frequency-swept interference, the time interval during which each sweep is perturbed is little dependent upon the value of the first threshold. Figs. 4, 7, and 8 directly give the relative time that the radar set is disturbed. Comparative curves for the optimum-integer value of m are given in Fig. 9 as functions of the mean duty cycle p_i of the interference. It will, for instance, be seen that for a moderately intense interference with $p_i=0.2$, without coincidence the radar receiver is disturbed for 3.5 per cent of the time, whereas with double and triple coincidence the figure is lowered to 0.05 per cent and 0.005 per cent, respectively. With a highly intense interference of $p_i=0.5$, which in fact would imply that an ordinary PPI is completely dis-

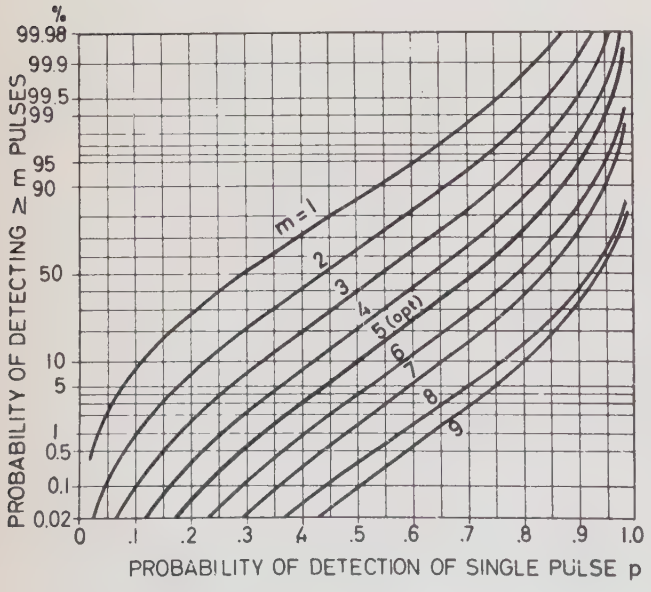

Fig. 7—Probability of detection in a system employing double coincidence techniques. $n=10$.

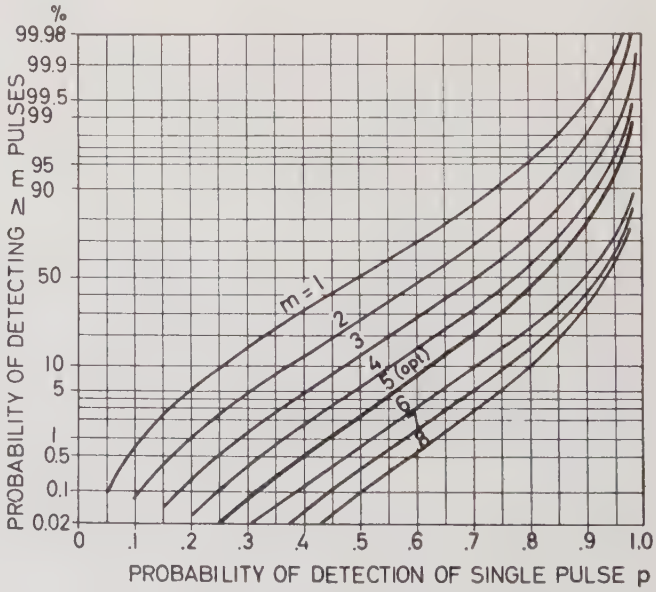

Fig. 8—Probability of detection in a system employing triple coincidence techniques. $n=10$.

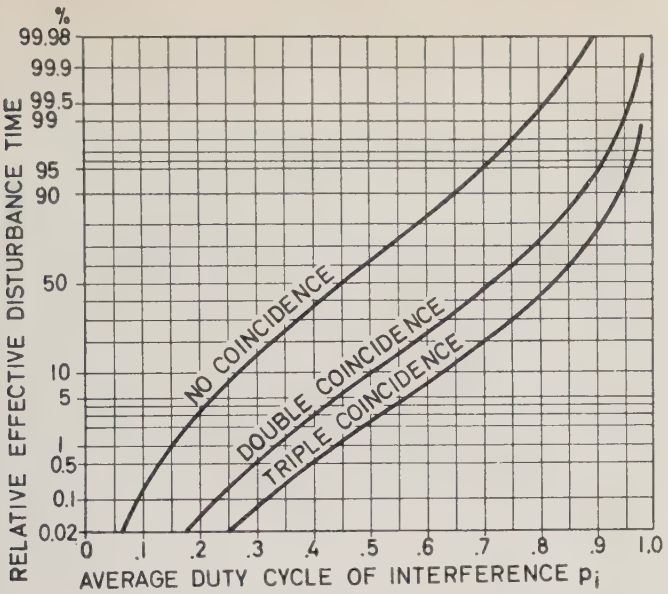

Fig. 9—Effective disturbance time. $n=10$. $m_{opt}=5$ in all three cases.

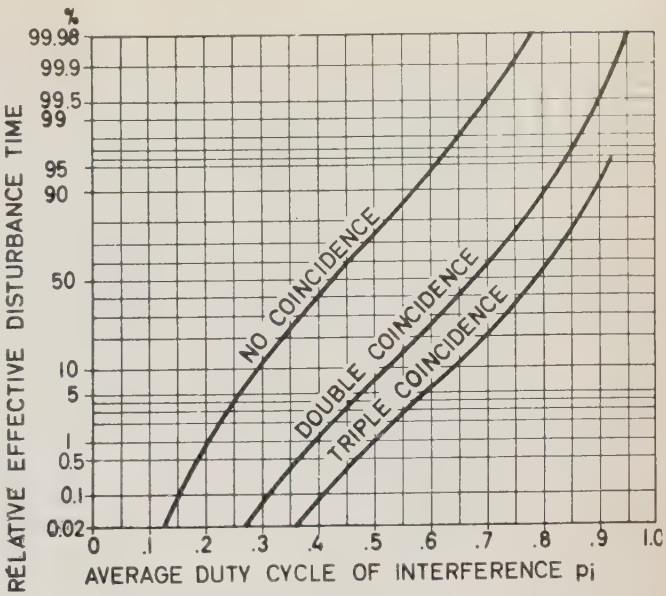

Fig. 10—Effective disturbance time. $n=20$. $m_{opt}=9$ in all three cases.

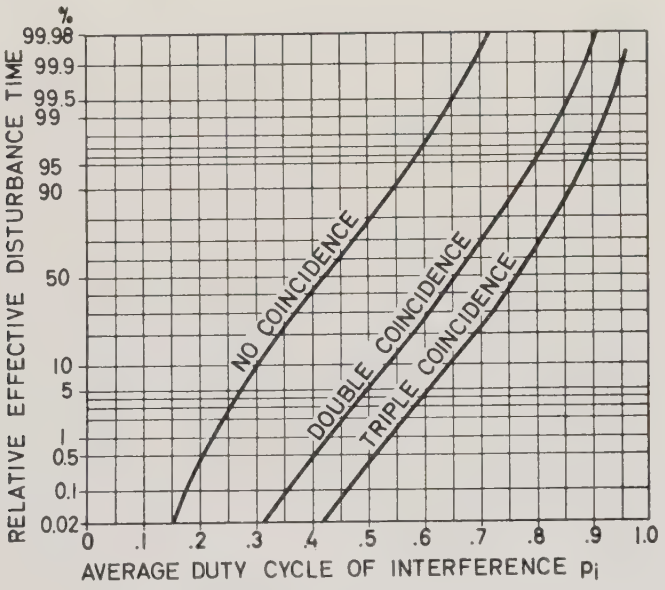

Fig. 11—Effective disturbance time. $n=30$. $m_{opt}=13$ in all three cases.

turbed, the figures are 62 per cent, 10 per cent, and 2.4 per cent, respectively.

Example 2. Noise Conditions ($n=10$)

Fig. 9 shows the advantage to be expected under the jamming types which the circuits are designed to combat. Next one might ask, what is the disadvantage, if any, under random (Gaussian) noise conditions? As an example, we may require that the false-alarm rate should equal 10^{-5}, and we seek the signal-to-noise ratios required to give a detection probability of 50 per cent or 90 per cent. Worked-out examples are given in Table I where the figures and equations used also are indicated.

It appears from Table I that for a steady signal, the required signal-to-noise ratio is increased by about 0.6 db when double diversity is used, and by 1.2 db for triple diversity, for both values of p_D indicated. For noise-like signal and $p_D=90$ per cent, double and triple coincidence require 3.2 db and 6.6 db better signal-to-noise ratio, respectively, than when no coincidence is used. For $p_D=50$ per cent, the corresponding figures are 2.3 db and 4.2 db.

An increase in required signal-to-noise ratio implies that the radar range is reduced as indicated in Table II, which assumes that the radar system obeys an inverse fourth-power law.

TABLE I

DETERMINATION OF REQUIRED SIGNAL-TO-NOISE RATIO
$n=10(m=5)\cdot p_{FA}=10^{-5}$

No. of coincidence inputs		One		Two		Three	
Probability p_D of detection		90 per cent	50 per cent	90 per cent	50 per cent	90%	50%
Probability p_0 that a noise sample exceeds R_t		0.033 [Eq. (3)]		0.112 [Eq. (6)]		0.16 [Eq. (8)]	
Relative threshold $R_t/\sqrt{N_0}$		2.61 (Fig. 1 or 2 ($-\infty$ db))		2.12 (Fig. 1 or 2 ($-\infty$ db))		1.92 (Fig. 1 or 2 ($-\infty$))	
Probability p that signal-plus-noise sample exceeds R_t		0.65	0.45 (Fig. 4)	0.864	0.71 (Fig. 7)	0.94	0.82 (Fig. 8)
Required signal-to-noise ratio S_0/N_0 db	Steady signal	6	4 (Fig. 1)	6.6	4.6 (Fig. 1)	7.2	5.2 (Fig. 1)
	Noise-like signal	8.4	5 (Fig. 2)	11.6	7.3 (Fig. 2)	15	9.2 (Fig. 2)

TABLE II

RELATION BETWEEN REQUIRED INCREASE IN SIGNAL-TO-NOISE RATIO AND RANGE

Reduction in range, per cent	6	12	16	21	25	29	35
Increase in signal-to-noise ratio, db	1	2	3	4	5	6	7

CONCLUSION

When coincidence techniques are used in a double-threshold detection system, a great advantage is obtained when the interference consists of pulses of low effective duty cycle. The advantage is considerable even for high duty cycles which would in effect completely block a noncoincidence system. The advantage is greater the higher the number of coincidence inputs. Against suitably synchronized interference the advantage is lost.

When the interfering signal is random noise, coincidence techniques reduce the range. The reduction is greater the larger the number of coincidence inputs. Due to the form of the radar equation, this disadvantage may be unimportant as compared with the advantage obtained under random-pulse interference conditions.

APPENDIX I

SYSTEM WITHOUT COINCIDENCE

Consider Fig. 3(a). The probability that a video pulse exceeds the first threshold is denoted by p, and the number of samples is n. The probability of detection, *i.e.*, of receiving at least m pulses from the counter output, is found by adding terms of the binomial distribution,

$$p_D = \sum_{x=m}^{n} \binom{n}{x} p^x (1-p)^{n-x}. \tag{1}$$

This may also be expressed as

$$p_D = \binom{n}{m} p^m + (-1)^m \sum_{x=m+1}^{n} (-p)^x \binom{n}{x}\binom{x-1}{m-1} \tag{2}$$

If p_D is *very small*, as may be the case when only noise is present, the first term in (2) predominates, and the remaining terms have alternating signs. The probability of false alarm therefore is given by

$$p_{FA} \approx \binom{n}{m} p_0^m , \tag{3}$$

where p_0 is the probability that a noise pulse exceeds the first threshold.

APPENDIX II

COINCIDENCE SYSTEM WITH TWO INPUTS

Consider Fig. 3(b). The n samples are delayed by one sweep period and correlated with the undelayed signal, as shown in the example below, where digit 1 corresponds to a pulse out from the first threshold device

Input	1	0	1	1	1	0	1	1	0	0
Delayed input		1	0	1	1	1	0	1	1	0	0
Output	0	0	1	1	0	0	1	0	0	

The coincidence circuit will give an output only if the delayed and undelayed sample are simultaneously 1. This requires the input signal to contain at least two 1's

in close configuration. More generally, $r+1$ 1's in close configuration will give out r 1's.

The probability of getting k 1's in a predetermined order in the input signal is

$$q(k) = p^k(1-p)^{n-k}. \tag{4}$$

The number of permutations containing k input 1's and x output 1's is denoted $P(k, x)$, and the probability of detection is given by

$$p_D = \sum_{x=m}^{n-1} \sum_{k=x+1}^{k_{\max}} p^k(1-p)^{n-k}P(k,x) + \Delta(x)\cdot(1-p)^n. \tag{5a}$$

In (5a), $\Delta(x)=1$ for $x=0$ and is zero otherwise. By a rather cumbersome counting of permutations, one finally arrives at[5]

$$P(k, x) = \binom{n+1-k}{k-x} \cdot\left[1 + \sum_{r=1}^{x}\binom{k-x-2+r}{r}\right], \tag{5b}$$

and

$$k_{\max} = \tfrac{1}{2}[n + x + \tfrac{1}{2}(1-(-1)^{n+x})]. \tag{5c}$$

When the probability $p=p_0$ that a noise pulse exceeds the first threshold is *very low*, (5) simplifies to

$$p_{FA} \approx p_0{}^{m+1}(n-m). \tag{6}$$

Appendix III

Coincidence System with Three Inputs

For the case illustrated in Fig. 3(c), the output will contain a 1 only if there are three 1's in close configuration in the input signal, *e.g.*,

Input	0 1 1 0 1 0 1 1 1 0 0
Delayed T	0 1 1 0 1 0 1 1 1 0 0
Delayed $2T$	0 1 1 0 1 0 1 1 1 0 0
Output	0 0 0 0 0 0 1 0 0

[5] K. Endresen and R. Hedemark, "Video Sweep to Sweep Coincidence Techniques for Radar Receivers Employing a Double-Threshold Mode of Detection," Norwegian Defence Res. Establishment, Internal Rept. No. T-209; November, 1960.

The derivation proceeds along the same pattern as in Appendix II. One first determines the number of favorable permutations $P(k, x)$ containing k input and x output 1's. Next is calculated the maximum value of k. Omitting the details, one finds after a laborious procedure[5]

$$p_D = \sum_{x=m}^{n-2} \sum_{k=x+2}^{k_{\max}} p^k(1-p)^{n-k}P(k,x) + \Delta(x)R(p,n) \tag{7a}$$

where

$$P(k, x) = \sum_{r=0}^{g(n,k,x)} \binom{(n+1-k)}{0,\ 5(k-x+N)+r} \cdot\binom{0,\ 5(k-x+N)+r}{N+2r}, \cdot\left[1 + \sum_{s=1}^{x}\binom{0,\ 5(k-x-N)-2-r+s}{s}\right], \tag{7b}$$

$$N = \tfrac{1}{2}[1-(-1)^{k-x}], \tag{7c}$$

$$\left.\begin{aligned} g(n,k,x) &= \tfrac{1}{2}(k-x-2-N) \quad \text{for } n \geq 2k-x-2 \\ g(n,k,x) &= \tfrac{1}{2}(2n-3k+x+2-N) \quad \text{for } n < 2k-x-2 \end{aligned}\right\}, \tag{7d}$$

$$k_{\max} = \tfrac{1}{3}\{x + 2[n+1-\alpha(n,x)\} + g(\alpha), \tag{7e}$$

$$\alpha(n,x) = 0, 1 \text{ or } 2, \text{ chosen to make } k_{\max} \text{ an integer}, \tag{7f}$$

$$g(\alpha) = \begin{cases} 1 & \text{for } \alpha(n,x) = 2 \\ 0 & \text{for } \alpha(n,x) \neq 2 \end{cases}, \tag{7g}$$

$$\Delta(x) = \begin{cases} 1 & \text{for } x = 0 \\ 0 & \text{for } x \neq 0, \end{cases} \tag{7h}$$

$$R(p,n) = \sum_{k=0}^{u}\binom{n+1-k}{k} p^k(1-p)^{n-k}, \tag{7i}$$

$$u = \tfrac{1}{4}[2n+1-(-1)^n]. \tag{7j}$$

This exact equation, which is well suited for digital computation, should be used for all values of p_D greater than of the order $10^{-3}-10^{-4}$. For very small values of p_D, (7) simplifies into

$$p_{FA} \approx p_0{}^{m+2}(n-m-1). \tag{8}$$

Correspondence

Antenna and Receiving-System Noise-Temperature Calculation*

A Naval Research Laboratory report[1] recently completed by the writer contains results, which are here summarized, that may be of interest.

In Part I a calculated curve representing the noise temperature of a typical directive antenna in the frequency range 100 to 10,000 Mc is presented (Fig. 1) together with the method and details of the calculation. Since antenna noise temperature (averaged over all galactic directions) is virtually independent of antenna gain and beamwidth, the curve may be used as an approximation for any typical directive antenna. The assumed environmental conditions are: 1) average cosmic noise; 2) solar noise temperature ten times the quiet level, with the sun in a unity-gain minor lobe of the antenna pattern; 3) a cool-weather temperate-zone atmosphere; and 4) a fixed ground-noise contribution of 36°K, as would result with a ground at blackbody temperature 290°K viewed over a π-steradian solid angle by minor lobes averaging 0.5 gain (3 db below isotropic level). The values given by this curve may readily be modified for other conditions.

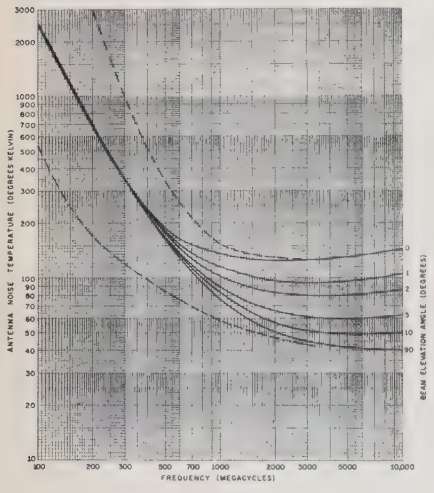

Fig. 1—Antenna noise temperature for typical conditions. Dashed curves are for maximum and minimum cosmic and atmospheric noise.

Part II presents a methodology for utilizing this antenna noise temperature in calculation of a system noise temperature, from which the total system noise power output and signal-to-noise power ratio may easily be computed. Basic concepts and definitions are first reviewed and then applied to development of formulas for noise temperatures of components of a cascade system and of the over-all system. The need for definition of both spot (frequency-dependent) noise temperature and average temperature over a pass band is pointed out, together with the need for definition of a transducer input noise temperature that reflects only the intrinsic noise.

The IRE-defined input noise temperature[2] of a two-port transducer, T_e, is interpreted to include, in the case of a multiple-input-response transducer, the effect of noise power contributed by a standard-temperature input termination via the spurious responses. For cascade-system noise-temperature calculation, a more suitable definition (in the case of a multiple-response heterodyne transducer) is obtained by postulating a zero-temperature (noise-free) input termination. The resultant transducer *output* noise temperature, divided by the conversion gain for a particular response, gives an input noise temperature referred to that response. The response thus chosen is usually designated the principal response, and this temperature is therefore called the principal-response input noise temperature, T_p. It is equivalent to T_e with the noise contribution of the standard-temperature input termination via the spurious responses deleted. Its relation to the transducer noise factor F is

$$T_p = (F - \beta)T_0, \tag{1}$$

where $T_0 = 290°K$ and β is defined by the relation

$$\beta = \frac{\int_0^\infty G(f)df}{\int_p G(f)df}. \tag{2}$$

$G(f)$ is the available transducer conversion power gain as a function of input frequency, and the integral of the denominator is taken over the principal-response band only. Since by definition,

$$T_e = (F - 1)T_0, \tag{3}$$

it follows that

$$T_p = T_e - (\beta - 1)T_0. \tag{4}$$

In a cascade system the noise-temperature contribution of a broad-band component that precedes a multiple-response transducer is shown to be expressible, subject to some mild conditions, as the sum of the spot-temperature values in the various response channels weighted by the relative gains of the channels. Utilizing this result, a formula for the noise temperature of a basic system consisting of antenna, transmission line, and a multiple-response receiver having s responses is obtained:

$$T_N = T_p + \sum_{i=1}^{s} [G(f_i)/G(f_p)] \cdot [T_a(f_i)/L_r(f_i) + T_r(f_i)]. \tag{5}$$

The frequencies f_i are the nominal frequencies of the response bands, and f_p is the principal-response frequency. L_r is the available-loss factor of the transmission-line system, including antenna dissipative losses, and T_r is the effective transmission-line output noise temperature, given by

$$T_r = T_t(1 - 1/L_r), \tag{6}$$

where T_t is the thermodynamic (Kelvin) temperature of the transmission-line lossy material. For the case of a single-response transducer ($\beta = 1$), (5) reduces to

$$T_N = T_a/L_r + T_r + T_e, \tag{7}$$

where the quantities on the right-hand side are evaluated at the nominal response frequency. (This equation, together with Fig. 1, appeared in a recent paper on radar range calculation.)[3]

This system noise temperature is referred to the receiver input terminals. If the effective signal power at this point is P_r, the receiver output signal-to-noise power ratio is

$$S/N = P_r/(kT_NB_N), \tag{8}$$

where k is Boltzmann's constant and B_N is the over-all system noise bandwidth. By suitable definition of P_r this result is applicable to simultaneous reception of signal in more than one response of a multiple-response system, as sometimes occurs in radiometry and possibly in other applications.

The system noise temperature may also be referred to the system input terminals. It is pointed out that this reference point is the only one for which a system noise temperature can be used to compare the low-noise merit of different systems, and then only if standard environmental conditions (*e.g.*, those of Fig. 1) are postulated, and if the system "input terminals" are defined to precede *all* dissipative losses, including any in what would ordinarily be considered the antenna (*i.e.*, such losses must be included in evaluation of the transmission-line loss factor, L_r). The expressions for the system noise temperature referred to the system input, corresponding to (5) and (7), are

$$T_{NI} = L_r(f_p)T_p + \sum_{i=1}^{s} [G(f_i)/G(f_p)] \cdot [T_a(f_i) + L_r(f_i)T_r(f_i)] \tag{9}$$

and

$$T_{NI} = T_a + L_r(T_r + T_e). \tag{10}$$

At the National Symposium of the IRE PGMTT in May, 1961, while the NRL report was in preparation, a five-member panel presented results of a study[4] of the

* Received by the IRE, July 7, 1961.

[1] L. V. Blake, "Antenna and Receiving-System Noise-Temperature Calculation," Naval Res. Lab., Washington, D. C., NRL Rept. No. 5668; 1961.

[2] IRE Standard 59 IRE 20.S1, "Methods of measuring noise in linear twoports, 1959" (Proc. IRE, vol. 48, pp. 60–68; January, 1960. See p. 68.)

[3] L. V. Blake, "Recent advancements in basic radar range calculation technique," IRE Trans. on Military Electronics, vol. MIL-5, pp. 154–164; April, 1961.

[4] H. A. Haus, *et al.*, "Elementary considerations of noise performance," *The Digest of Technical Papers*, 1961 PGMTT Natl. Symp., pp. 53–57.

noise-temperature method of analyzing system noise performance. They propose characterizing a multiple-response transducer by a "broad-band effective input noise temperature," T_b, which is related to T_p by the formula

$$T_p = \beta T_b. \tag{11}$$

Therefore, system noise temperature is expressible in terms of either of these quantities through this relation. Another difference in the approach of the PGMTT panel is their use of signal bandwidth, rather than system noise bandwidth, in definition of system noise temperature. Through consideration of both approaches it is hoped that more comprehensive IRE standards on transducer and system noise temperature may eventuate.

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Nonsymmetrical Properties of Nonlinear Elements in Low- and High-Impedance Circuits*

The interest to date in parametric devices has primarily been concerned with the power conversion arising from the first-order nonlinear term in a reactance expansion. For this case it makes no difference whether a particular nonlinear element is used in series or parallel resonant circuits since the power conversion is independent of the circuit configuration. However, when the higher-order nonlinear terms are to be used for power conversion or when the perturbations on first-order power conversion due to the higher-order nonlinearities are to be considered, the circuit configuration which terminates the nonlinear device has an important influence on the operating characteristics.

To illustrate this point consider the two common methods of reactively terminating a nonlinear capacitor in order to avoid power conversion into unwanted harmonic frequencies and mixing products. The capacitor may look into a low impedance at these frequencies so that a large number of harmonic currents flow without significant voltage components, or the termination may be a high-impedance one so that a large number of harmonic voltages exist without significant currents. These two cases are illustrated by the amplifier circuits of Fig. 1, in which the voltage across the capacitor in Fig. 1(a) and the current through the capacitor in Fig. 1(b) consist essentially of the sum of only three components.

To analyze these circuits two expansions of the nonlinearity of the capacitance are useful. For a semiconductor capacitor the usual approximation for the capacitance is

* Received by the IRE, July 13, 1961.

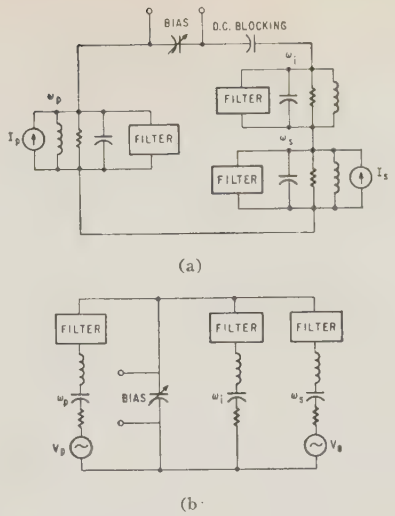

Fig. 1—Parametric amplifier circuits. (a) Low-impedance termination. (b) High-impedance termination.

$$C(v) = \frac{dq}{dv} = \frac{C_0}{\left[1 - \frac{V_B + v}{\phi}\right]^n} = \left[\frac{C_0\phi^n}{(\phi - V_B)^n}\right] \cdot \frac{1}{\left[1 - \frac{v}{\phi - V_B}\right]^n} = \frac{C_B}{\left[1 - \frac{v}{\phi - V_B}\right]^n}, \tag{1}$$

where C_B is the average capacitance at the (generally negative) bias voltage V_B, and v and q are the ac variations about the bias point. The charge variation on the capacitor may be expanded as a Taylor series in the voltage as

$$q(t) = C_B v(t) + \mathcal{C}_a v^2(t) + \mathcal{C}_b v^3(t) + \cdots = C_B v(t) + \frac{nC_B v^2(t)}{2!(\phi - V_B)} + \frac{n(n+1)C_B v^3(t)}{3!(\phi - V_B)^2} + \frac{n(n+1)(n+2)C_B v^4(t)}{4!(\phi - V_B)^3} + \cdots \tag{2}$$

and the voltage in terms of the charge as,

$$v(t) = S_B q(t) - \mathcal{S}_a q^2(t) - \mathcal{S}_b q^3(t) - \cdots = \frac{1}{C_B} q(t) - \frac{nq^2(t)}{2!(\phi - V_B)C_B^2} - \frac{(1-2n)nq^3(t)}{3!(\phi - V_B)^2 C_B^3} - \frac{(2-3n)(1-2n)nq^4(t)}{4!(\phi - V_B)^3 C_B^4} - \cdots \tag{3}$$

Although either of these expansions may be used in the analysis of these circuits, the use of the improper expansion is mathematically cumbersome and does not lead to a good physical understanding of the characteristics of the circuit. Thus, the expansion should be made in terms of the variable consisting of the sum of the fewest components, with (2) being used for the circuit of Fig. 1(a), and (3) for Fig. 1(b).

In the three frequency systems illustrated, the even terms in the equations cause power conversion. In this respect the two circuits are equivalent, if higher-order perturbations are neglected, as may be seen by computing the figure of merit, often used, of relative change in the average capacitance or elastance due to the pumping.

$$\frac{\Delta C_B}{C_B} = \frac{\mathcal{C}_a \left| v(t) \right|}{C_B} = \frac{n \left| v(t) \right|}{2!(\phi - V_B)} \tag{4}$$

$$\frac{\Delta S_B}{S_B} = \frac{\mathcal{S}_a \left| q(t) \right|}{S_B} = \frac{n \left| q(t) \right|}{2!(\phi - V_B)C_B} = \frac{n \left| v(t) \right|}{2!(\phi - V_B)}. \tag{5}$$

The other terms in the expansions do not have this symmetry. The third term leads to a susceptance in each resonant circuit of Fig. 1(a) with the value

$$jB_i = j\tfrac{3}{4}\mathcal{C}_b(\left| V_i \right|^2 + 2\left| V_k \right|^2 + 2\left| V_l \right|^2), \tag{6}$$

and likewise a reactance in the circuit of Fig. 1(b) of

$$jX_i = j\frac{3}{4}\frac{\mathcal{S}_b}{\omega_i^2} \cdot \left(\frac{\left| I_i \right|^2}{\omega_i^2} + 2\frac{\left| I_k \right|^2}{\omega_k^2} + 2\frac{\left| I_l \right|^2}{\omega_l^2}\right), \tag{7}$$

where the V's and I's are the peak values of the voltages across and the currents through the varactor. These terms cause a signal-dependent and mutually-coupled susceptance or reactance which produces nonsymmetrical resonance curves and accounts for the jump phenomenon observable in nonlinear resonant systems. The significance of these expressions is that the reactance perturbation may be eliminated by using an abrupt-junction varactor ($n = \frac{1}{2}$) while the same cannot be done for the susceptances with presently available semiconductor capacitors. When the third term in the voltage expansion is eliminated by using an abrupt-junction varactor, all higher-order terms which cause reactive and power conversion perturbations also vanish. This result predicts that an abrupt-junction varactor used in a high-impedance circuit will give low phase distortion under high signal conditions, and that this is a good method of building large signal amplifiers and phase-distortionless limiters. This has been supported experimentally by measurements of phase distortion in parametric limiters using abrupt and graded-junction varactors at 30 Mc.[1]

In situations where higher-order terms in the expansions are to be used for power conversion, such as in parametric amplifiers using LF pumping, in third and higher harmonic generators, and in one-third and lower subharmonic generators, the series-resonant configuration should not be used with an abrupt-junction varactor.

A plot of the rate at which the first three nonlinear terms in the voltage expansion change with the junction exponent is shown in Fig. 2. It can be seen that the $\mathcal{S}_b$ term goes through 0 at $n = \frac{1}{2}$ with a rather steep slope so that to achieve good cancellation of this term careful selection of the varactor will be necessary, and biasing of the device should

[1] Private communication with W. T. Jones of Sylvania Waltham Labs., Waltham, Mass.

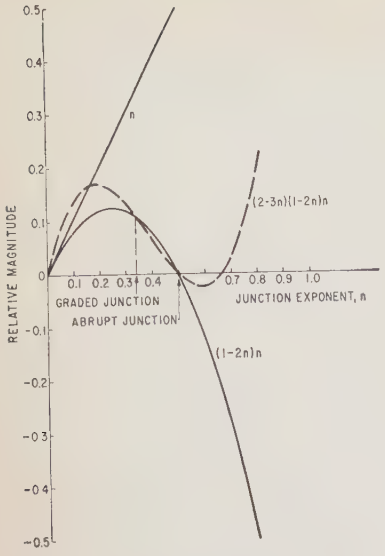

Fig. 2—Relative magnitude of terms in the voltage expansion.

be done so that (1) is valid over the expected range of voltage variation with constant n. It is also clear that $n=\frac{2}{3}$ is a good exponent value to use in systems depending upon the third term in the expansion for power conversion since this will eliminate all higher-order perturbations. However, the reactance perturbation corresponding to (7) will still be present in this case.

Further developments in semiconductor technology may make it possible to control the individual nonlinear terms in the charge and voltage expansions and therefore to make possible tailor-made nonlinear elements for given applications. Until then, however, we must be content to use the available nonlinear characteristics as judiciously as we can.

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Improved Tungsten Evaporation Filament for Gold-Silver Alloy*

In the fabrication of Mesa transistors, base stripes are formed by evaporating a gold-silver alloy in a vacuum of less than 5×10^{-5} mm of Hg. The stripe thickness, 2000 Å, is very critical and plays a major role in the electrical characteristics of the finished units. In order to control such thickness to a high degree of accuracy, it is essential that the required amount of alloy be evaporated *completely*.

The evaporation takes place at the bottom of a tungsten conical basket filament, whose temperature is electrically controlled. A phenomenon, frequently arising in this process, is that the molten alloy does not adhere to the tungsten filament and drops off, thus making the complete evaporation impossible.

A solution to this wetting problem has been found by precoating the W filament with platinum. One inch of platinum wire, 0.020 inch in diameter, was coiled around the bottom of the conical basket and melted thereon. The resulting coating proved ideal as a medium to which the molten alloy adheres perfectly.

In order to ascertain that no significant amount of platinum was being evaporated in the process, a coated tungsten filament was weighed before and after 10 evaporation runs. Its weight change was compared with that of an uncoated control filament used for an equal number of runs. No appreciable difference in weight change was found between the two filaments.

To determine that no deleterious but undetected amount of platinum was being deposited on the stripes and affecting the electrical properties, 1000 devices processed with the new technique and 1000 control devices were compared. No significant difference was found between the two groups.

Platinum-coated tungsten filaments, which last as long as the uncoated filaments, have completely eliminated the lack of adhesion problem, thus offering considerable savings in time, labor and expensive materials.

These results extend the findings of L. Holland,[1] who remarks that Ag can be kept in W basket by binding fine platinum wire on the outside.

The author is grateful to Dr. R. J. Gnaedinger, Jr. for valuable comments.

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* Received by the IRE, August 1, 1961.

[1] L. Holland, "Vacuum Deposition of Thin Films," Chapman and Hall, Ltd., London, Eng., p. 112; 1958.

Some Operating Characteristics of Flash-Pumped Ruby Lasers*

In the course of a series of experiments with ruby optical masers, several observations have been made which may be useful in the planning and interpretation of future experiments.

Five lasers of various characteristics were used. All of them were made according to conventional designs.[1] Evaporated silver coatings were applied so that one end had an estimated transmission of about 1 per cent and the other end was opaque. Pumping was performed by a GE Type 524 helical quartz, Xenon-filled photoflash lamp. All experiments were done at room temperature. The pumping energy was supplied by an appropriate electronic circuit and a condenser bank capable of reaching the maximum ratings of the flash lamp (4000 volts and 4000 joules).

It was found that the pumping energy required to start laser oscillation depends upon the dimensions of the laser and upon the condition of its surfaces. Table I gives some threshold voltage values interpolated from the measurements. Fig. 1 shows curves of measured threshold voltages V for the various lasers. Two curves each for lasers Nos. 1, 4, and 5 show the effect of resilvering.

The thresholds could be determined easily and repeatably to within ±5 volts. Placement of the ruby within the flash tube could be varied by $\frac{1}{4}$ in without changing the threshold voltage more than 50 volts.

One unexpected result was that the pumping energy alone did not characterize adequately the laser threshold, contrary to the assumption used in previous publications.[1,2] Fig. 1 shows that the threshold pumping energy W and charge Q vary in a linear manner with variations of the supply capacitance C, for a given laser and flash tube. It appears that the highest efficiency is obtained with high voltages and an appropriately small condenser.

To explore the nature of the "capacitance effect" and to explain the straight lines plotted in Fig. 1, preliminary measurements of the light emitted by the flash tube and by a laser were made. For each measurement, a Type 931A photomultiplier was used, driving an oscilloscope through a coaxial cable properly terminated so as to preserve the HF structure of the light pulses.

To sample the flash-tube light, a diffuser of ground sapphire rod was mounted at the center of the flash-tube helix. The light was filtered through blue-green copper chelate solution, which transmits from about 4200 to 6200 Å, in an attempt to eliminate light at wavelengths not directly useful for pumping, and was attenuated by several layers of paper. The laser beam was also attenuated with paper. Both attenuators were adjusted until the largest signals obtained were within the almost-linear range of the photomultipliers.

The pumping light pulses were found to have a smooth exponential rise lasting about 0.2 msec, one or two sharp peaks, and a smooth exponential fall which was 90 per cent completed in 1.7 msec for 558 mfd and 0.8 msec for 258 mfd. The laser beam began after a delay depending upon the voltage applied. The envelope of its oscillogram was about the same shape, and ceased at about the same time, as the pumping light pulse. The envelope has a superimposed band of "noise" pulses presumably arising from statistical fluctuations of the gain of the laser oscillator, resembling those reported elsewhere. The frequencies present exceeded 100 kc and appeared to increase with the voltage.

Table II shows that the pumping light pulse has a duration t nearly proportional to the capacitance C used, and an amplitude h governed chiefly by the applied voltage V,

* Received by the IRE, June 26, 1961; revised manuscript received, July 31, 1961.

[1] R. J. Collins, *et al.*, "Coherence narrowing directionality and relaxation oscillation in the light emission from ruby," *Phys. Rev. Lett.*, vol. 5, no. 7, pp. 303–305; October, 1960.

[2] M. Ciftan, *et al.*, "A ruby laser with an elliptic configuration," Proc. IRE, pt. 1, vol. 49, pp. 960–961; May, 1961.

TABLE I
CHARACTERISTICS OF SEVERAL RUBY LASERS

No.	Diam. (in)	Length (in)	C-Axis orientation	Threshold with 300 mfd		
				Potential (volts)	Charge (coulombs)	Energy (joules)
1	0.500	1.37	0°	3500	1.05	1840
2	0.250	1.50	0°	3000	0.90	1350
3	0.250	1.50	0°	2900	0.87	1260
4	0.100	0.75	0°	2610	0.78	1020
5	0.100	0.75	90°	2550	0.76	970

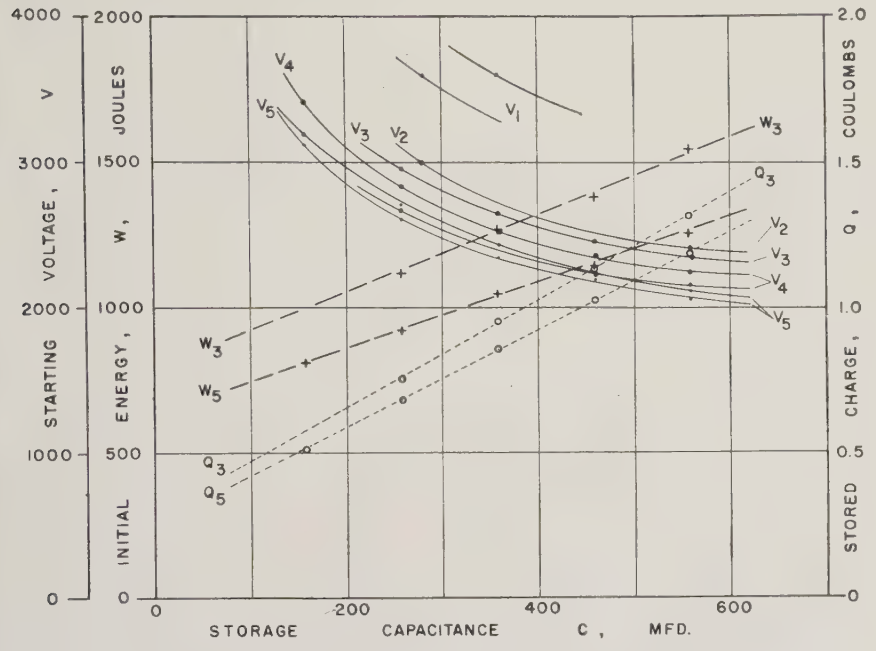

Fig. 1—Threshold conditions for five different lasers. (Two different silverings used on some.)

TABLE II
LIGHT PULSE CHARACTERISTICS

Xenon Lamp, (Blue-Green)					No. 5 Laser (Red)		
Capacitance mfd	Initial voltage	Peak amplitude volts	Peak time, msec	Duration msec	Peak amplitude volts	Start time, msec	Duration msec
558	4000	0.15	0.2–0.3	3.5	0.16	0.1	1.7
	3500	—	—	—	—	—	—
	3000	0.07	0.2,0.4	3.5	0.04	0.3	1.0
	2500	—	—	—	0.0075	0.6	0.7
	2000	0.015	0.2–0.6	3.5	0.000	—	—
258	4000	0.14	0.2–0.25	1.5	0.10	0.1	0.7
	3500	—	—	—	0.06	0.25	0.65
	3000	0.07	0.2,0.3	1.5	0.01	0.4	0.4
	2500	—	—	—	—	—	—
	2000	0.017	0.2	2.0	0.000	—	—

approximately as $h=KV^n$. This fact, together with the capacitance effect, gives the formula $h=(A/t+B)^n$, implying that the minimum light intensity required to start laser action is an inverse function of the time over which it is applied, plus a constant.

A simple test with selected polaroids which were effective in the red part of the spectrum showed that the stimulated optical emission from laser No. 5 was plane polarized. The electric vector of the emerging laser beam was at right angles to the optic axis of the crystal, in agreement with a result recently reported by others.[3] Wide variations of pumping energy did not change the effect. The much weaker fluorescent output, however, did not seem to be plane polarized. As expected, no trace of plane polarization could be found in the beams from laser No. 3 and laser No. 4, which are optically symmetrical about their cylindrical axes. Attempts to find circularly polarized light failed also.

The author wishes to thank the following for their valuable suggestions, help, and encouragement during this work: Dr. R. J. Collins, Dr. G. Pish, G. Damewood, and W. Bradshaw.

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[3] I. D. Abella and H. Z. Cummins, "Thermal tuning of ruby optical maser," *J. Appl. Phys.*, vol. 32, pp. 1177–1178; June, 1961.

Repetitive Hair-Trigger Mode of Optical Maser Operation*

The ruby optical maser has a characteristic relaxation time between the $\bar{E}$ excited state and the ground state of the order of several milliseconds,[1] by means of spontaneous emission. This emission time is of course greatly reduced when the population inversion is sufficient to give rise to stimulated emission, *i.e.*, optical maser action.[2,3] For three-level devices such as ruby a very sizeable portion of the pumping energy is required just to transport half of the ground state population to the excited state and so bring the optical maser to the point of oscillation. The additional amount to provide stimulated emission is relatively small.

One may then conjecture that it would be useful to provide enough excitation to bring an optical maser to the almost oscillating condition, and, since there is then no stimulated emission and the relaxation time is relatively long, one can provide a relatively small amount of energy at a later time (provided this time is small compared to the relaxation time) and thus accurately control the time of the output with little or no jitter and achieve a kind of *hair-trigger* type of operation.

In the past control of the pumping pulse has been inadequate to obtain even repeatable performance let alone this hair-trigger operation. We have demonstrated an alternative method of obtaining a short pulse of coherent light. The idea was to use the indication of optical maser output as a kind of preparation state and then, by the application of a much less energetic sharp pulse some time later (short compared to the relaxation time), to stimulate a short emission. Since the optical maser crystal is just a shade below the threshold of oscillation, this relatively low-energy excitation pulse was found sufficient to bring about an immediate emission pulse.

A crude experiment was performed in which a ruby, encircled by a small spiral flash tube, is in turn placed within a second larger spiral flash tube so that the ruby and flash tubes are all coaxial. (See Fig. 1.) A pulse of light, fairly rectangular in shape, is caused to be emitted from the outer flash tube. This pulse has just sufficient duration for the ruby to begin emitting in optical maser action. Then the pulse is switched off, and the optical maser action ceases. Approximately 100 μsec later a very short fairly low-energy pulse is applied to the inner flash tube, and the ruby is seen to emit 1 or 2 μsec later. The emitted optical maser pulse is found to be much more vigorous than the pulse normally emitted, such as in the preparation state.

Fig. 2 shows the trace of such an experiment. Because of the unfavorable geometry of the outer flash tube, the *preparation-state*

* Received by the IRE, July 14, 1961; revised manuscript received, July 26, 1961.

[1] S. Sugano and Y. Tanabe, "Absorption spectra of Cr^{3+} in Al_2O_3," *J. Phys. Soc. (Japan)*, vol. 13, pp. 880–899; August, 1958.

[2] T. H. Maiman, "Optical maser action in Ruby," *Brit. Commun. and Electronics*, vol. 7, pp. 674–675; September, 1960. Also *Nature*, vol. 187, pp. 493–494; August, 1960.

[3] R. J. Collins, *et al.*, "Coherence, narrowing directionality and relaxation oscillations in the light emission from Ruby," *Phys. Rev. Lett.*, vol. 5, pp. 303–305; October, 1960.

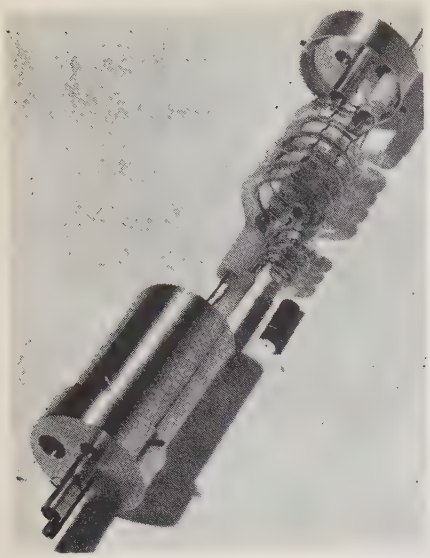

Fig. 1—Exploded view of coaxial flash tube and ruby used in hair-trigger LASER experiment.

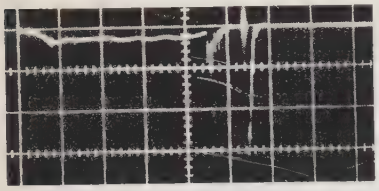

Fig. 2—Trace of LASER emission during hair-trigger experiment. Scale is 100 μsec per major division.

pulse was of 400 μsec duration and required approximately 1200 watt-seconds before emission occurred. The pulse from the inner flash tube was only 60 watt-seconds and yielded, as seen, an optical maser emission in the hair-trigger mode several times as large as obtained in the preparation state.

For applications such as the COLIDAR (*CO*herent *LI*ght *D*etecting *A*nd *R*anging) pulsed ranging system,[4] it would be highly desirable to have periodic, controllable low-jitter pulses.

We propose to accomplish this by use of the radiation of optical maser output as a preparation for what we shall call the *repetitive hair-trigger mode* of operation. That is, the optical maser would be excited to where it emits. The excitation would then be turned off, and short bursts of excitation applied with a periodicity short compared to the relaxation time. This would cause a periodic optical maser output in predictable fashion.

We make the simplified assumption that after the preparation phase the population dips slightly below the threshold and then decays exponentially. The energy required to excite a pulse in the repetitive hair-trigger mode is then the amount required to bring the system back up to threshold.

Let E_p = threshold energy for preparation phase, r = pulse repetition rate, E = threshold energy per pulse in the repetitive hair-trigger mode, p = the required power for the repetitive hair-trigger mode, ΔE_p = the immediate energy drop below threshold following an emission, and τ = characteristic decay time. Then roughly,

$$E \cong E_p(1 - \exp(-)1/r\tau) + \Delta E_p. \quad (1)$$

For $r \gg 1/\tau$ we obtain

$$E \cong E_p(1/r\tau) + \Delta E_p \quad (2)$$

and

$$p = rE \cong E_p/\tau + \Delta E_p. \quad (3)$$

Although precise values of the parameters in the above equations await experimental determination, it is evident that the much higher relative efficiency of this mode should make it possible to increase not only the duty cycle, but the peak output as well. It is obvious that the jitter between optical maser output and pump input should be greatly reduced.

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[4] M. L. Stitch, E. J. Woodbury, and J. H. Morse, "Optical ranging system uses laser transmitter," *Electronics*, vol. 34, pp. 51–53; April 21, 1961.
M. L. Stitch, F. J. Meyers, J. H. Morse, and E. J. Woodbury, "Breadboard COLIDAR (coherent light detecting and ranging) system (uncl.)," *Proc. 5th Natl. Convention on Military Electronics*, pp. 279–284; June, 1961.
D. A. Buddenhagen, B. A. Lengyel, F. J. McClung, Jr., and G. F. Smith, "An experimental laser radar," 1961 IRE International Convention Record, pt. 5, pp. 285–290.

WWV and WWVH Standard Frequency and Time Transmissions*

The frequencies of the National Bureau of Standards radio stations WWV and WWVH are kept in agreement with respect to each other and have been maintained as constant as possible with respect to an improved United States Frequency Standard (USFS) since December 1, 1957.

The nominal broadcast frequencies should for the purpose of highly accurate scientific measurements, or of establishing high uniformity among frequencies, or for removing unavoidable variations in the broadcast frequencies, be corrected to the value of the USFS, as indicated in the table below. The corrections reported have been arrived at by means of improved measurement methods based on LF and VLF transmissions.

The characteristics of the USFS and its relation to time scales such as ET and UT2 have been described in a previous issue,[1] to which the reader is referred for a complete discussion.

The WWV and WWVH time signals are also kept in agreement with each other. Also they are locked to the nominal frequency of the transmissions and consequently may depart continuously from UT2. Corrections are determined and published by the U. S. Naval Observatory. The broadcast signals are maintained in close agreement with UT2 by properly offsetting the broadcast frequency from the USFS at the beginning of each year when necessary. This new system was commenced on January 1, 1960. A retardation time adjustment of 20 msec was made on December 16, 1959; another retardation adjustment of 5 msec was made at 0 UT on January 1, 1961.

WWV Frequency
With Respect to U. S. Frequency Standard

1961 July	Parts in 10^{10}†
1	−149.0
2	−149.2
3	−149.1
4	−149.0
5	−148.9
6	−149.5
7	−149.6
8	−149.8
9	−150.0
10	−149.9
11‡	−149.8
12	−150.7
13	−150.4
14	−150.1
15	−149.8
16	−149.4
17	−149.5
18	−149.5
19	−149.7
20	−149.9
21	−150.0
22	−150.2
23	−150.4
24	−150.5
25	−150.9
26	−151.1
27	−151.2
28	−151.2
29	−151.3
30	−151.1
31	−150.9

† A minus sign indicates that the broadcast frequency was low. The uncertainty associated with these values is $\pm 5 \times 10^{-11}$.
‡ WWV adjusted -1×10^{-10} on July 11.

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* Received by the IRE, August 24, 1961.
[1] Refer to "National standards of time and frequency in the United States," Proc. IRE (Correspondence), vol. 48, pp. 105–106; January, 1960.

Correction to "Stable Low-Noise Tunnel-Diode Frequency Converter"*

In the second paragraph of the above,[1] the statement is made that any noise sources in the nonlinear resistance are completely mismatched in the case $g_c = g_0$, and $g_0/g_c \to 0$. This statement must be qualified: the noise sources will be completely mismatched (and the noise figure approach unity) only if in addition $Y_0/Y_1 \to 1$, where Y_0 and $2Y_1$ are, respectively, the average and the fundamental coefficients of the Fourier expansion of the equivalent modulated mean-square noise current in the nonlinear resistor.[2]

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* Received by the IRE, August 1, 1961.
[1] F. Sterzer and A. Presser, Proc. IRE (Correspondence), vol. 49, p. 1318; August, 1961.
[2] See, for example, M. J. O. Strutt, "Noise-figure reduction in mixer stages," Proc. IRE, vol. 34, pp. 942–950; December, 1956.

Lossy Coupling in Parametric Amplifier*

Heffner and Wade's method of analysis,[1] originally applied to a two-resonant circuit parametric amplifier with lossless coupling branch, may be extended successfully in the case of an amplifier with lossy coupling branch. Adopting their assumptions that 1) the network is linear, 2) voltages of frequencies other than resonant frequency may be neglected in respect to the resonant-frequency voltage across each circuit, and 3) the diode-cartridge capacitance may be included in parallel into both the resonant circuits (since it influences only resonant frequencies of the circuits), one comes to the equivalent circuit of the amplifier shown in Fig. 1. For simplicity, the pump and bias circuits are not indicated in the figure and the load resistance has been included in the loss-resistance of the first (amplifying action) or second (mixing action) resonant circuit. Furthermore, the cosinusoidal form of all the time-varying quantities is assumed in the following according to the general relation

$$a = a(\omega_a t) = \overline{A}\cos(\omega_a t + \phi_a) = \tfrac{1}{2}(Ae^{j\omega_a t} + A^* e^{-j\omega_a t}). \quad (1)$$

Thus, for the case $\omega_3 = \omega_1 + \omega_2$, one may calculate the current i_c in the coupling lossy branch according to

$$r_s i_c + \frac{1}{C_0 + c(\omega_3 t)}\int i_c dt = u_1(\omega_1 t) - u_2(\omega_2 t). \quad (2)$$

As the equation is linear, it may be split into two separate equations with exciting functions $u_1(\omega_1 t)$ and $-u_2(\omega_2 t)$, respectively, leading to two solutions i_{c1} and i_{c2}, each of which has components at angular frequencies ω_1 and ω_2. Using the perturbational method for solving (2) and taking advantage of earlier assumed linearity of the network which in the case of parametric diode means $C_0 \gg \overline{C}$, one may confine himself to consider only the first two approximation terms in the series solution.[2] Then, according to the general notation of (1), the complex amplitudes of the current i_c can be found as

$$I_{c1}' = \frac{U_1}{Z_c(\omega_1)};\quad I_{c1}'' = \frac{-CU_1^*}{2j\omega_1 C_0^2 Z_c^*(\omega_1) Z_c(\omega_2)}$$
$$I_{c2}'' = \frac{-U_2}{Z_c(\omega_2)};\quad I_{c2}' = \frac{-CU_2^*}{2j\omega_2 C_0^2 Z_c(\omega_1) Z_c^*(\omega_2)}, \quad (3)$$

where primes and double primes denote that complex amplitudes correspond to angular frequencies ω_1 and ω_2, respectively, and

$$Z_c(\omega) = r_s(\omega) + \frac{1}{j\omega C_0} \quad (4)$$

is the time-average impedance of the coupling branch of Fig. 1 at the given angular frequency ω.

* Received by the IRE, May 29, 1961; revised manuscript received, June 5, 1961.

[1] G. Wade and H. Heffner, "Gain, bandwidth and noise characteristics of a variable parameter amplifier," *J. Appl. Phys.*, vol. 29, pp. 1321–1331; September, 1958.

[2] K. Grabowski, Thesis, Dept. of Telecommunication, Gdansk Inst. Tech., Poland; 1960 (in Polish). Also, "An application of the perturbational method to the analysis of a parametric amplifier and mixer," *Bull. Acad. Polon. Sci.*, Sér. sci. techn., vol. 7, pp. 515–524; September, 1960 (in English).

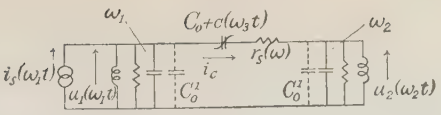

Fig. 1.

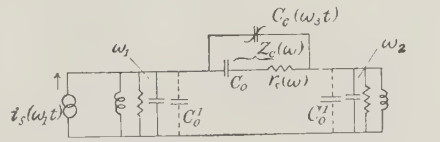

Fig. 2.

It is possible to show[2] that (3) leads to the equivalent circuit of the parametric amplifier shown in Fig. 2, where the amplitude of the equivalent variable capacitance $\overline{C}_e$ is related to the amplitude of the actual variable capacitance $\overline{C}$ according to the relation

$$\overline{C}_e = \overline{C}\{[1+\tan^2\delta(\omega_1)][1+\tan^2\delta(\omega_2)]\}^{-1/2}, \quad (5)$$

where

$$\tan\delta(\omega) = \omega C_0 r_s(\omega). \quad (6)$$

Now, the constant in time branch $Z_c(\omega)$ [from (4)] may be inserted in parallel in both the circuits, and finally, there remains a circuit which is identical to that of Heffner and Wade. Therefore, one may use all their formulas, replacing in them actual capacitance-amplitude $\overline{C}$ by $\overline{C}_e$ and adding to the admittances of both circuits the admittance $1/Z_c(\omega)$.

As a simple illustration of this procedure, let us examine the condition for the setting up of oscillations in the amplifier. For the lossless coupling, the condition[1] was

$$\overline{C}^2 \geq \frac{4G_1(\omega_1)G_2(\omega_2)}{\omega_1\omega_2}. \quad (7)$$

Now, for coupling with losses, by (5) and (7), there results

$$\overline{C}^2 \geq \frac{4G_{1c}(\omega_1)G_{2c}(\omega_2)}{\omega_1\omega_2}[1+\tan^2\delta(\omega_1)]\cdot[1+\tan^2\delta(\omega_2)], \quad (8)$$

where the subscript c at resonant conductances indicates that they include losses of the coupling branch.

In the case of ideal circuits, *i.e.*, those in which the only losses in the network are those of the series resistance of the diode, the condition (8) becomes

$$C^2 \geq 4C_0^2 \tan\delta(\omega_1)\tan\delta(\omega_2). \quad (9)$$

Introducing the capacitance modulation factor $m_c = \overline{C}/C_0$, the last formula may be rewritten in a more pronounced manner as

$$[\tan(\omega_1)\tan(\omega_2)]^{1/2} \leq \frac{m_c}{2}. \quad (10)$$

In the microwave range, the parameter $r_s(\omega)$ usually does not depend on frequency; therefore, the condition for the setting up of oscillations in the amplifier finally becomes

$$r_s C_0 \omega_{av} \leq \frac{m_c}{2}, \quad (11)$$

where ω_{av} is the geometric mean value of resonant angular frequencies of the amplifier.

Since, here, both resonant circuits have been eliminated by assuming them lossless, (10) or (11) might be useful as a definition of cutoff frequencies of a given parametric diode with an available capacitance-modulation factor m_c, average bias capacitance C_0 and series resistanec r_s.

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Discussion of Calculation of False Alarm Rate*

In their interesting article, Thaler and Meltzer[1] give an expression for false alarm rate, which is

$$F = M\frac{Q(T)}{Q(O)}, \quad (1)$$

where F is the false alarm rate, M is the expected number of positive crossings of the mean, $Q(T)$ is the probability density of the noise amplitude at threshold setting T (where T is in units of standard deviations measured from the mean), and $Q(O)$ is the probability density of the noise amplitude at the mean. They state that this expression holds whenever the threshold is set sufficiently far above the mean level of the noise.

Expression (1) does not actually hold in general. A more correct expression is

$$F = M\frac{\int_T^\infty Q(x)dx}{\int_0^\infty Q(x)dx}, \quad (2)$$

where $Q(x)$ is the probability density of the noise amplitude, and x is the amplitude in units of standard deviations measured from the mean.

In one of the cases analyzed in the paper, that of the square-law detector, (1) and (2) do agree. The agreement occurs in this case, however, only because the post-detector probability density function happens to be a simple negative exponential, the chi-squared function with two degrees of freedom.

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* Received by the IRE, February 28, 1961.

[1] S. Thaler and S. Meltzer, "The amplitude distribution and false alarm rate of noise after post-detection filtering," Proc. IRE, vol. 49, pp. 479–485; February, 1961.

Author's Comment[2]

The expression for false alarm rate given in our paper is an experimental result of the simulation study described. In each instance studied, the false alarm rate (defined as the number of threshold crossings per second) after post-detection filtering was found to be proportional to the amplitude

[2] Received by the IRE, March 20, 1961.

probability density, rather than to the integral of the density as suggested by Rogers. The former conclusion is the one expected since it agrees with a theoretical result (applicable to noise with a Gaussian amplitude distribution) given by Rice.[3] The integral of the probability density equals the product of the average duration of each threshold crossing and the frequency of such crossings. Since both change in the same way with threshold level, the integral changes faster than the frequency of the crossings.

When the threshold is too close to the mean level of the noise, some of the threshold crossings become so close together that it may not be appropriate to count each crossing as a separate false alarm. This effect was negligible at useful false alarm rates. The elimination of some threshold crossings at very high false alarm rates would have caused the false alarm rate to change more slowly with threshold than the frequency of threshold crossings.

S. Thaler
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[3] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. J.*, vol. 23, pp. 282–332, July, 1944; vol. 24, pp. 46–156, January, 1945.

Microwave Radiation Hazards*

Mumford's paper[1] offers a much needed summary of the biological hazards of RF radiation, plus data on prediction, measurement and instrumentation. Going one step further, we have also observed that in addition to the biological hazards of direct exposure to microwave radiation, there can be hazardous secondary effects to technicians working in field intensities that might be considered acceptable from a biological standpoint.

Objects made of conducting materials that are located in field intensities of only one or two milliwatts per square centimeter can easily serve to couple sufficient RF energy to be hazardous. For example, observations have been made where a man carrying a wrench in each hand formed a dipole, coupling sufficient energy to produce a warming sensation in the body. It was also observed that an electrical shock can be produced when two wrenches or two rods are brought together, and then separated drawing a small arc. In the particular incident, the radiation sources was a pulse radar which produced a fairly substantial peak power. The voltage causing the shock was undoubtedly the envelope of the RF resulting from arc rectification. The electrical shock received from field intensities of only one or two milliwatts per square centimeter was sufficient to discourage venturing into a higher field intensity to repeat the experiment.

While it appears improbable that an electrical shock hazard exists from the standpoint of an electrocution, this condition should be given recognition. Obviously a man working in such a precarious position as on a steel tower might have sufficient reaction to such a shock as to cause him to fall. On the basis of these experiences it would appear advisable to show due respect for field intensities even as low as one-tenth milliwatt per square centimeter.

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* Received by the IRE, March 2, 1961.

[1] W. W. Mumford, "Some technical aspects of microwave radiation hazards," Proc. IRE, vol. 49, pp. 427–447; February, 1961.

Power Limiting in the 4-kMc to 7-kMc Frequency Range Using Lithium Ferrite*

Extensive investigations into the use of yttrium iron garnet (YIG) for microwave power limiters have been reported in the literature.[1–4] The threshold for limiting occurs at particularly low power levels if the microwave frequency and biasing dc magnetic fie d are chosen for *coincidence* of the uniform precession and subsidiary ferromagnetic resonances. The coincidence region of a sphere is the frequency octave defined by $4\pi M/3 \leq \omega/\gamma \leq 8\pi M/3$, where M is the saturation magnetization and γ is the gyromagnetic ratio. For a YIG sphere with $4\pi M = 1750$ gauss, the coincidence region extends from 1600 Mc to 3200 Mc. Nonmagnetic ions such as Ga and Al can be substituted into YIG to lower the magnetization[5] and to shift the coincidence region to lower frequencies. *Coincidence* limiters have been made to operate in the 1000-Mc region with Ga-YIG.

The purpose of this note is to report experimental evidence indicating the practicability of extending the useful range of coincidence limiters to about 7000 Mc. The coincidence octave for a sphere of lithium ferrite[6] which has a $4\pi M$ of 3900 Gauss at room temperature, extends from 3640 Mc to 7280 Mc covering a frequency range of particular importance in microwave communications. The limiter structure consists of two rectangular cross-section, half-wavelength strip transmission-line cavities, and its operation is essentially the same as those previously employed.[3] It will suffice here to present the limiter characteristics.

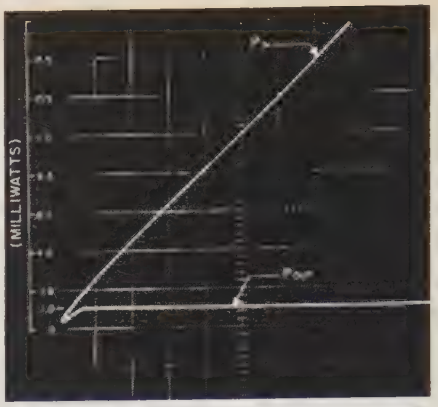

Fig. 1—Characteristic of the lithium ferrite limiter at 5200 Mc. The upper curve shows the input power and the lower curve shows the limited output power. At 1 w of input power the output remained less than 1.5 mw.

Limiting was observed at 5200 Mc at an output power level of 1 mw with an essentially flat characteristic which rose only to 1.5 mw at about 1 w of input power. Fig. 1 is a picture of an oscilloscope presentation of the output power vs input power showing the sharp break at the limiting threshold. The limiting power level can be raised to other predetermined values by simple modifications of the microwave structure. The single crystal sphere used in this limiter had $\Delta H = 8$ oersteds and $\Delta H_k = 3.6$ oersteds for spin waves of 5000 Mc measured along the [111] (easy) axis[7] of magnetization.

In view of the present extensive use of the 4000-Mc to 7000-Mc frequency range, the results presented here indicate that the *coincidence* limiter using lithium ferrite may find important applications in microwave systems.

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Murray Hill, N. J.

* Received by the IRE, August 1, 1961.

[1] R. W. DeGrasse, "Low-loss gyromagnetic coupling through single crystal garnets," *J. Appl. Phys.*, vol. 30, pp. 155–156S; April, 1959.

[2] C. N. Patel, "Investigations of Magnetically Tunable Narrow Bandpass Nonreciprocal Filters Using Ferrimagnetic Resonators," Stanford Electronics Labs., Stanford, Calif., Tech. Rept. no. 411-1; April, 1961.

[3] F. J. Sansalone and E. G. Spencer, "Low-temperature microwave power limiter," IRE Trans. on Microwave Theory and Techniques, vol. MTT-9, pp. 272–273; May, 1961.

[4] P. S. Carter, "Magnetically tunable microwave filters using single-crystal yttrium iron garnet resonators," IRE Trans. on Microwave Theory and Techniques, vol. MTT-9, pp. 252–260; May, 1961.

[5] E. G. Spencer and R. C. LeCraw, "Linewidth narrowing in gallium substituted yttrium iron garnet," *Bull. Am. Phys. Soc.*, pt. 1, ser. 2, vol. 5, p. 58; January 1960.

[6] The lithium ferrite samples used in this work were grown by J. W. Nielsen at Bell Telephone Labs., and with considerable improvement later at Airtron, Inc.

[7] R. T. Denton and E. G. Spencer, "Ferromagnetic resonance losses in lithium ferrite as a function of temperature," *J. Appl. Phys.* (to be published). See also the earlier paper by: A. D. Schnitzler, V. J. Folen, and G. T. Rado, "Ionic ordering effects in the ferromagnetic resonance of lithium ferrite monocrystals," *J. Appl. Phys.*, vol. 31, pp. 348–349S; May, 1960.

Phase Inverters Utilizing Controlled Superconductors*

In a recent article[1] it was pointed out that controlled superconductors may be employed in several classes of linear amplifiers. One amplifier configuration discussed

* Received by the IRE, May 26, 1961.

[1] P. M. Chirlian and V. A. Marsocci, "The Controlled superconductor as a linear amplifier," IRE Trans. on Component Parts, vol. CP-8, pp. 84–88; June, 1961.

was a push-pull circuit such as the one shown in Fig. 1. Such push-pull circuits may be operated class B to improve their operating efficiency. It is the purpose of this note to show that the phase-inverting driver transformer is not necessary since the phase inversion may be implemented directly by the use of controlled superconductors. Therefore, the space-saving aspects of the thin-film geometry can be maintained. Two examples of phase-inverting circuitry are shown in Fig. 2(a) and (b). It is of particular interest to note that in the circuit of Fig. 2(a) the phase inversion is obtained by the interesting property of the controlled superconductor which is that the phase of the output signal may be shifted 180° by reversing the polarity of one of the bias batteries.

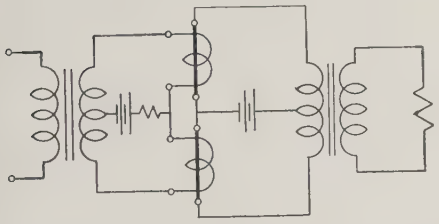
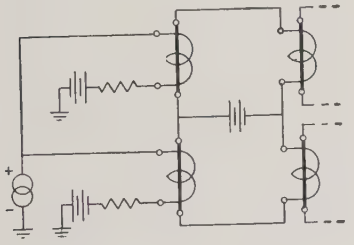
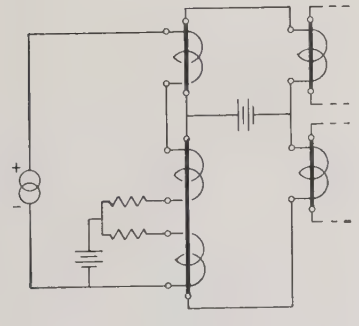


Fig. 1—A controlled-superconductor push-pull amplifier. Note that the symbolism used here applies to all types of controlled superconductors and not just to wire cryotrons.



(a)



(b)

Fig. 2—Phase inverters employing controlled superconductors.

This property is further discussed in the above-mentioned article. The second phase-inverter circuit only requires one input-bias battery; however, it has the disadvantage that it requires an additional control element.

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A Ferrimagnetic Limiter-Isolator*

It has been generally known for some time that a ferrimagnetic power limiter can be built.[1,2] When the RF power level in a ferrimagnetic material exceeds a certain threshold value, and given a particular set of physical circumstances, a subsidiary absorption peak will appear at a value of dc field somewhat below that required for resonance. This nonlinear characteristic[3] of ferrites and garnets can be utilized to build a limiter-isolator.

The threshold power level for the onset of nonlinear absorption is very important in determining the usefulness of a microwave power limiter. In practice this critical level is determined not only by the intrinsic properties of the material and the frequency of operation, but also by the size and shape of the material and the structural configuration in which the material is used. In rectangular waveguide this threshold power level has been studied using a vertical slab of nickel-zinc ferrite in the configuration shown in Fig. 1.

The critical level for nonlinear absorption can be attained at a lower incident power level for a small slab of ferrimagnetic material if it is placed in the region of circular polarization as shown from the data in Fig. 1. Care must be taken to insure that the slab does not appreciably effect the normal microwave fields in the waveguide. This region of circular polarization of the microwave magnetic field is also noted as the region utilized in designing ferrimagnetic isolators. With the material located in this region, the "spin waves" (the mechanism for nonlinear absorption) receive little or no excitation from the forward wave in the waveguide; however, the reverse wave heavily excites the spin waves and hence absorption is found to be nonreciprocal.

Data taken at X band with the material in dielectrically loaded double ridge waveguide are shown in Fig. 2. The material in this case was yttrium-aluminum-iron garnet. The limiting action is noted as the difference between absorption at low power and at 10 kw for the reverse wave. The nonreciprocal action of the structure is seen by comparing the forward wave with the reverse wave data at high power. Similar data are shown in Fig. 3 where yttrium-iron garnet slabs are located near the region of circular polarization in double ridge C-band waveguide. The threshold in this structure is seen to be approximately 100 watts. The small absorption peak in the forward wave at the field for main resonance results from locating the material slightly out of the region of circular polarization. Lower thresholds have been obtained; however, it has been found more difficult to obtain good circular polarization when using these low threshold configurations.

* Received by the IRE, June 26, 1961.

[1] G. S. Uebele, "Characteristics of ferrite microwave limiters," IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 18–23; January, 1959.

[2] R. L. Martin, "High-power effects in ferrite slabs at X-band," *J. Appl. Phys.*, vol. 30, pp. 1595–1605; April, 1959.

[3] H. Suhl, "The theory of ferromagnetic resonance at high signal powers," *J. Phys. and Chem. of Solids*, vol. 1, pp. 209–227; April, 1957.

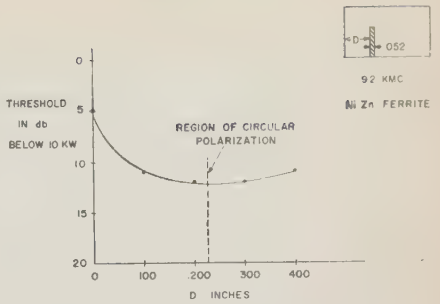

Fig. 1—Threshold power level (db below 10 kw) for a nickel-zinc ferrite as a function of position in rectangular X-band waveguide.

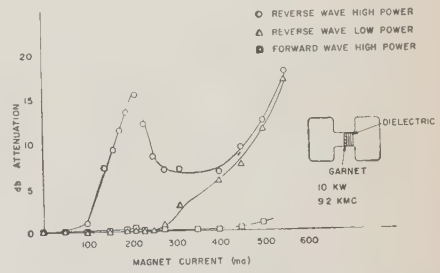

Fig. 2—Subsidiary resonance region showing the high and low power response of an yttrium-aluminum-iron garnet in dielectrically loaded double ridge X-band waveguide.

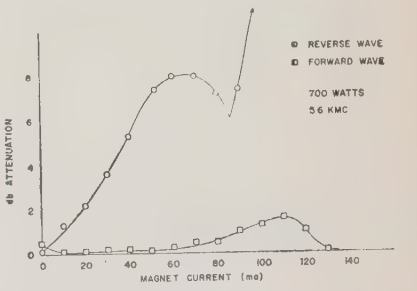

Fig. 3—Subsidiary resonance region using yttrium-iron garnet slabs near the region of circular polarization in double ridge C-band waveguide.

If the ferrimagnetic material is sufficiently long, the attenuation above the threshold will increase with increasing power, resulting in a constant back load on the microwave source. Thus, high power isolators can be constructed utilizing the subsidiary resonance. A device operating in this fashion can also be used as a power limiter, as illustrated in Fig. 2 by the low insertion loss of the low power reverse wave. Possibly the most unique application of the device would be the use of it in the antenna arm of the duplexer of a radar system. It would pass the high power transmitted pulse virtually unattenuated, but would attenuate any high power reflections from the antenna, or high power interfering signals from other sources and would pass any low power received signals virtually unattenuated. It would thus be an isolator, a limiter, and a high power interference suppressor. All of these functions are performed passively and simultaneously.

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On the Impedance of Long Wire Suspended Over the Ground*

Much has been written over the years on the subject of the impedance of wires lying on the ground or suspended above it.[1] The possibility that a long horizontal wire will be a feasible radiator of VLF radio waves has reopened interest in the problem. It is the purpose of this note to outline a rather simple solution for the impedance of an infinitely long wire located at a height h over a homogeneous flat ground.

Choosing a rectangular coordinate system (x, y, z), the ground is defined by the space $z<0$. The wire is located at $z=h$ and is parallel to the x axis. Now for VLFs (*i.e.*, less than 30 kc), and since the height h is to be much less than the wavelength, it is permissible to neglect displacement currents both in the air and in the ground. For a uniform current I in the wire of radius a, the electric field has only an x component and is given formally by[2,3]

$$E_x = \frac{i\mu\omega}{2\pi} I \left[\log \left\{ \frac{x^2 + (h+z)^2}{x^2 + (h-z)^2} \right\}^{1/2} + 2\int_0^\infty \frac{e^{-\lambda(h+z)}}{\lambda + (\lambda^2+\gamma^2)^{1/2}} \cos(\lambda x) d\lambda \right],$$

where $\gamma^2 = i\sigma\mu\omega$ and μ is the magnetic permeability (assumed constant everywhere) and σ is the conductivity of the ground. It follows from the basic definition that the self impedance Z per unit length is given by

$$Z = \lim_{\substack{z\to h \\ x\to a}} \left[\frac{E_x}{I} \right].$$

Therefore

$$Z = \frac{i\mu\omega}{2\pi} \left[\frac{1}{2} \log \left(1 + \frac{4h^2}{a^2} \right) + P \right],$$

where

$$P = 2\int_0^\infty \frac{e^{-2\lambda h}}{\lambda + (\lambda^2+\gamma^2)^{1/2}} \cos \lambda a \, d\lambda.$$

This can be rewritten as

$$P = P_+ + P_-,$$

where

$$P_\pm = \int_0^\infty \frac{e^{-\lambda(2h \pm ia)}}{\lambda + (\lambda^2+\gamma^2)^{1/2}} d\lambda.$$

Introducing the variable $g=\lambda/\gamma$ and the parameter $\alpha = \gamma(2h \pm ia)$, the above is written as

$$P_\pm = \int_0^{\infty e^{-i\pi/4}} \frac{e^{-\alpha g}}{g + (1+g^2)^{1/2}} dg.$$

The integration contour now extends from the origin along a straight line in the fourth quadrant which makes an angle of 45° with the real axis. Noting that $\alpha \cong |\alpha| e^{-i\pi/4}$ and that the poles of the integral are of the imaginary axis, it is seen that the integration contour can be deformed to the real axis. Then, multiplying numerator and denominator by $(1+g^2)^{1/2}-g$, it readily follows that

$$P_\pm = \int_0^\infty [(1+g^2)^{1/2} - g] e^{-\alpha g} dg = T(\alpha) - \frac{1}{\alpha^2},$$

where

$$T(\alpha) = \int_0^\infty (1+g^2)^{1/2} e^{-\alpha g} dg.$$

The definite integral $T(\alpha)$ can be expressed in terms of known functions as follows:

$$T(\alpha) = \frac{\pi}{2\alpha} [H_1(\alpha) - Y_1(\alpha)],$$

where $H_1(\alpha)$ is the Struve's function[4] of order one, and $Y_1(\alpha)$ is the Bessel function of the second type of order one. The final solution is thus,

$$P = T[\gamma(2h+ia)] - \frac{1}{\gamma^2(2h+ia)^2} + T[\gamma(2h-ia)] - \frac{1}{\gamma^2(2h-ia)^2}.$$

For purposes of computation it is necessary to employ the series formula,

$$H_1(\alpha) = \frac{2}{\pi} \left[\frac{\alpha^2}{1^2\cdot 3} - \frac{\alpha^4}{1^2\cdot 3^2\cdot 5} + \frac{\alpha^6}{1^2\cdot 3^2\cdot 5^2\cdot 7} - \cdots \right],$$

and tables for $Y_1(\alpha)$ which are available for complex values of α.[5] For small values of α corresponding to low heights, the following formula may be useful:

$$T(\alpha) - \frac{1}{\alpha^2} = -\frac{\Gamma + \log(\alpha/2)}{2} + \frac{1}{4} + \frac{\alpha}{3} + \text{terms in } \alpha^2, \alpha^3, \text{ etc.}$$

Γ is Euler's constant and is equal to 0.5773.

Of particular interest is the real part of the impedance, or resistance ΔR, per unit length. For the usual case $h \gg \alpha$ and thus

$$\Delta R \cong \frac{\mu\omega}{8} \left[1 - \mathrm{Re}\left(i\frac{16}{3}(\gamma h) \right) + \text{terms in } (\gamma h)^2, (\gamma h)^3, \text{ etc.} \right] \cong \frac{\mu\omega}{8} \left[1 - \frac{8\sqrt{2}}{3} \sqrt{\sigma\mu\omega}\, h + \cdots \right].$$

Or if $\sqrt{\sigma\mu\omega}\cdot h \ll 1$,

$$\Delta R \cong \frac{\mu\omega}{8},$$

which is conveniently written

$\Delta R = \pi^2/10 f_{kc}$ ohms per km when f_{kc} is the frequency in kc/sec.

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* Received by the IRE, May 22, 1961.

[1] E. D. Sunde, "Earth Conduction Effects in Transmission Systems," D. Van Nostrand Co., Inc., New York, N. Y.; 1949. (A good survey of the early work is given here.)

[2] J. R. Wait, "Fields of a line source of current over a stratified conductor," *Appl. Sci. Res.*, sec. B, vol. 3, pp. 279 292; 1953.

[3] J. R. Carson, "Wave propagation in overhead wires with ground return," *Bell Sys. Tech. J.*, vol. 5, pp. 539–554; October, 1926.

[4] G. N. Watson, "Theory of Bessel functions," Cambridge University Press, Cambridge, Eng. 2nd ed.; 1944. (Struve's function is defined on page 328.)

[5] Natl. Bur. of Standards, "Tables of Bessel Functions $Y_0(z)$ and $Y_1(z)$ for Complex Argument," Columbia University Press, New York, N. Y.; 1950.

Masers, Lasers, and the Ether Drift?*

It is the purpose of this letter to suggest that the ether drift experiment of Cedarholm, Bland, Havens, and Townes[1] (hereafter referred to as CBHT), performed with two opposed ammonia masers, has given a null result because it is not affected by an ether drift, contrary to the expectations of its authors. From a consideration of the error in this experiment a new ether drift test employing an optical laser is proposed.

Two of the authors[2] have apologized for performing the experiment, since it seemed obvious to every right-thinking physicist that the null result was inevitable. On the other hand, the experiment, if true, could have considerable impact, since it has moved the upper limit for an ether drift from the previous value of 1.5 km/sec established by Joos, down to 0.015 km/sec.[3] Since this is less than the motion of a terrestrial laboratory due to the earth's rotation (about 0.3 km/sec in temperate latitudes), it would seem fruitless to look any more for ether drift effects. However, because of the importance of determining just this point, namely, whether or not the velocity of light is constant with respect to the earth's gravitational field, it was felt desirable to examine closely the basis for the null result of the CBHT experiment.

The theory is best given in the authors' words,[1] discussing "the change in frequency of a beam-type maser due to ether drift, assuming the molecules in the beam to have a velocity u with respect to the cavity through which they pass, and the cavity to have a velocity v with respect to the ether. The shift may be simply discussed by assuming that if v is zero, radiation is emitted perpendicularly to the molecular velocity so that there is no Doppler shift. If the cavity and beam are then transported at velocity v through the ether in a direction parallel to u, radiation must be emitted by the molecules slightly forward at an angle $\phi = \pi/2 - v/c$ with respect to u. The fractional change in frequency due to the Doppler effect is then $E = u/c \cos\phi$ or uv/c^2 due to motion through the ether, assuming that the proper molecular frequencies are unchanged by such motion." It is further assumed that this Doppler-shifted radiation is received by the cavity walls, and hence becomes the operating frequency of the maser. Since the shift depends on the magnitude and orientation of the ether drift v with respect to u, two masers with beams going in opposite directions should show a change in beat frequency when their orientation is changed with respect to the earth, or a diurnal effect due to the earth's rotation if they are kept in a fixed position.

This theory was first propounded by Moeller[4] in a far from lucid fashion. The flaw in the theory arises from the fact that

* Received by the IRE, August 4, 1961.

[1] J. P. Cedarholm, et al., "New experimental test of special relativity," *Phys. Rev. Lett.*, vol. 1, p. 342; November, 1958.

[2] J. P. Cedarholm and C. H. Townes, "A new experimental test of special relativity," *Nature*, vol. 184, p. 1350; October 31, 1959.

[3] C. H. Townes, "Quantum Electronics," Columbia University Press, New York, N. Y., p. 581; 1960.

[4] C. Moeller, "On the possibility of terrestrial tests of the general theory of relativity," *Il Nuovo Cimento*, (Suppl.), vol. 6, ser. 10, no. 1, p. 381; 1957.

both Moeller and CBHT believe that the wavefronts from the molecule are slightly tilted with respect to the molecule's velocity by the ether drift. A close examination of Moeller[5] discloses no mechanism for producing a tilt, and leads to the conclusion that the wavefronts of radiation emitted at right angles to the molecule's velocity arrive at the cavity wall still parallel to the molecule's velocity and with no Doppler shift, regardless of ether drift.

Following a suggestion of Brillouin,[6] we may represent the radiation from a molecule by the idealized diagram of Fig. 1. Here the upper plane B, illuminated by plane-parallel radiation, is traveling with the beam velocity u relative to the stationary plane C, which will represent the cavity walls. The aperture in plane B allows radiation to reach C, and will represent a molecule which is emitting radiation at right angles to the beam velocity. The path of the radiation from the molecule to the cavity walls is then as shown. We see that, although to the cavity walls the radiation appears to be coming in a line inclined to the perpendicular by the angle $\theta=u/c$, the wavefronts themselves arrive unchanged in direction, having traveled the intervening space with velocity c. They arrive at C at the same rate with which they left B, and there is no Doppler shift in this case. If now we add an ether drift v to the left, as shown, the only effect of this will be to increase the angle of deflection of the beam to the value $\theta'=(u+v)/c$ while leaving the wavefronts still parallel to the beam and the cavity walls. Once again, these wavefronts still arrive at the cavity wall with velocity c, and no Doppler shift is produced.

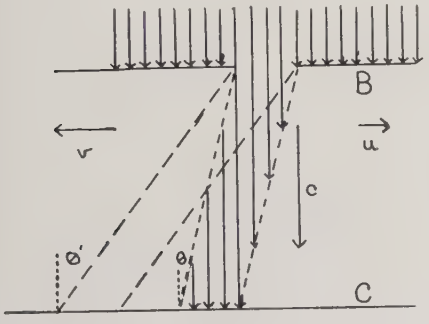

Fig. 1.

This is really just an elaborate way of explaining why, in prerelativity physics, no Doppler effect is found at a receiver when its distance from a transmitter remains constant, regardless of any ether drift. Since the transmitter is the ammonia molecule, and its distance from the cavity wall, the receiver, remains constant, no Doppler effect can be produced.

We see then that the null result of the ammonia maser experiment is a sort of built-in effect, inasmuch as the result will always be zero, regardless of any ether drift.

[5] C. Moeller, "The Theory of Relativity," Oxford University Press, New York, N. Y., ch. 1, pp. 1–15, 1957.

[6] L. Brillouin, "Wave Propagation and Group Velocity," Academic Press, Inc., New York, N. Y., p. 14; 1960.

From a consideration of Fig. 1, which was used by Brillouin to illustrate stellar aberration, one is led to inquire, "Why not perform a terrestrial aberration experiment to measure small ether drifts?" For a drift of 0.3 km/sec, which might be due to the earth's rotation, the aberration angle will be $0.3/3\times10^5$ or 10^{-6} radians. Since recent experiments by Jones[7] have measured the deflection of a light beam down to 2×10^{-12} radians, the aberration experiment should be easy. Alas, this is not the case, and for a curious reason. It has been pointed out to the author that, using conventional point light sources and collimators, a terrestrial aberration experiment for measuring ether drift is impossible. This is because the source and collimating lens are also immersed in the drift, if it exists, and a deflection of the beam beyond the lens will not be found, since the ether drift will cause inclined wavefronts to enter the collimator; the angle of inclination being just sufficient to compensate for the ether drift deflection beyond the lens.

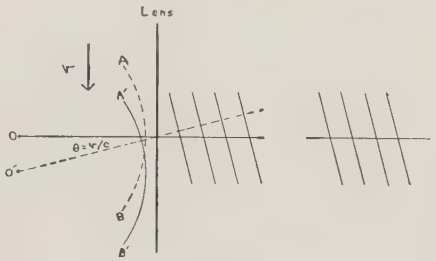

Fig. 2.

This can be seen from Fig. 2, which shows what happens in a collimator when a transverse ether drift exists. In the absence of an ether drift, the lens sees the spherical wavefront AB emitted from O, and turns this into a plane wavefront normal to the axis. In the presence of transverse drift, the lens sees the wavefront $A'B'$, which is displaced from AB in the direction of v. This wavefront appears to the lens to come from O', a point off the axis. From the geometry, a central ray through O' makes an angle $\theta=v/c$ with the axis, and the lens turns the displaced wavefront into an almost plane wavefront inclined to the axis by the same angle. The pencil enclosing these inclined waves continues to travel parallel to the axis, since the ether drift balances the tendency of the waves to depart from the direction of the axis. A few moments of thought will convince the reader that reflection from mirrors, taking account of the relative change in ether drift, will not in any case allow the reflected pencil of inclined wavefronts to depart from the path pursued by a similar pencil of normal wavefronts in the absence of ether drift.

This compensation is so good that one almost suspects that it is another law of nature that we cannot detect an ether drift by an aberration experiment. However, many of the effects observed in crystal optics are due to a similar anisotropy, and are observable only because a pencil of plane wavefronts normal to the pencil can be introduced into the crystal.

We may then look for a source of such wavefronts, not affected by ether drift, and a likely candidate seems to be the optical laser. This device is constrained to emit the largest part of its radiation with wavefronts parallel to the end plates, since this is the steady-state mode preferred by the laser. In the presence of a transverse ether drift, it is probably possible to set up a mode of inclined wavefronts similar to the emergent beam of Fig. 2, in which the waves are reflected from the end plates in such a manner that the back and forth propagation of energy is in the direction of the laser axis. In the case of the mode comprising waves parallel to the end plates, a transverse ether drift would cause the wavetrain to drift away from the axis. On the other hand, the active nature of the laser process may compensate this loss.

In any event, theory does not now tell us which mode the laser prefers, and it will be of interest to try one as a source in an aberration experiment. Following Jones' technique, it should be possible to form a coarse grating on the emitting plate, and let its image, without further collimation, fall on a similar obscuring grating with a radiation detector behind it. For a separation of one meter, the image of the first grating should be displaced $\pm10^{-4}$cm for a v/c of 10^{-6}.

While the maser, in the form of the CBHT experiment, seems to have failed us, the laser may save the day. The laser experiment is now under way.

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[7] R. V. Jones, "The velocity of light in a transverse magnetic field," *Proc. Roy. Soc.*, vol. 260, p. 47; February, 1961.

Matrix Analysis of Networks Having Infinite-Gain Operational Amplifiers*

It is the object of this note to describe and prove a method of analyzing linear-lumped time-invariant networks that contain an operational amplifier having infinite gain. The generalization for the case of several operational amplifiers in a network is immediate and need not be described.

An operational amplifier of gain A operating between nodes i and j introduces a constraint into the network by forcing the voltage e_j of node j (with respect to the reference node) to follow that of node e_i so as to maintain the relation

$$e_j = Ae_i \qquad (1)$$

without, however, loading node i. The method of analyzing constrained networks

* Received by the IRE, May 5, 1961.

is applicable to this case.[1] It prescribes in this instance the following way of constructing the required admittance matrix Y from the admittance matrix $\boldsymbol{Y}$ of the network without the constraint: in $\boldsymbol{Y}$ add column j multiplied by A to column i and subsequently delete row and column j. The result is Y. This procedure eliminates the constrained node j from the network equations.

The method is not immediately applicable to the case $A \to \infty$ because of the apparent singularity introduced by the transition to infinity. In this case of infinite gain we shall eliminate the source currents operating into node j, as before, but shall retain e_j and eliminate e_i, the voltage of the driving node. The following simple rule obtains: to obtain Y, delete in $\boldsymbol{Y}$ the row corresponding T to the driven node and the column corresponding to the driving node.

Turning now to the proof of this rule, we could start from the rule that applies to finite constraints. An independent proof will be found simpler and more instructive.

Eq. (1) becomes, for $A = \pm\infty$,

$$e_i = 0, \tag{2}$$

making node i a "virtual ground." The operational amplifier injects into node j a current $\boldsymbol{i}_{(s)j}$, of such intensity as to enforce (2). Shifting i and j for convenience into leading positions in the matrix representation, we have now (2) and

$$Y\begin{bmatrix} e_i \\ e_j \\ \vdots \end{bmatrix} = \begin{bmatrix} i_{(s)i} \\ i_{(s)j} \\ \vdots \end{bmatrix} = \begin{bmatrix} \\ +i_{(s)j} \\ \vdots \end{bmatrix} = i_{(s)}, \tag{3}$$

where the right-hand side is the source-current vector. Making use of (2) in (3) we see that column i in $\boldsymbol{Y}$ and the entry e_i can be omitted on the left-hand side. We intend to eliminate the unknown $\boldsymbol{i}_{(s)j}$ as well. This is achieved by deleting the jth equation of (3), *i.e.*, by deletion of row j in $\boldsymbol{Y}$ and entry j in $i_{(s)}$. The result is

$$Y\begin{bmatrix} e_j \\ \vdots \end{bmatrix} = \begin{bmatrix} i_{(s)i} \\ \vdots \end{bmatrix}. \tag{4}$$

Y is used, as always, for the calculation of system functions. For example, provided p is not driven and q is not driving,

$$H_{pq} = e_q/e_p = (Y^{-1})_{qp}i_{(s)p}/(Y^{-1})_{pp}i_{(s)p} = Y^{pq}/Y^{pp}, \tag{5}$$

where superscripts denote cofactors.

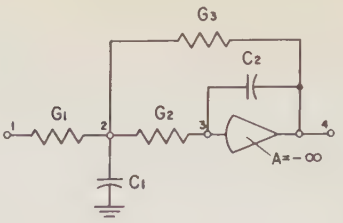

Fig. 1.

In general, Y can be written down without first writing $\boldsymbol{Y}$. As an example consider the "double integrator" of Fig. 1. We have

$$\boldsymbol{Y} = \begin{array}{c} \\ 1 \\ 2 \\ 3 \\ 4 \end{array}\begin{array}{c} \begin{array}{cccc} 1 & 2 & 3 & 4 \end{array} \\ \begin{bmatrix} G_1 & -G_1 & 0 & 0 \\ -G_1 & G_1+G_2+G_3+sC_1 & -G_2 & -G_3 \\ 0 & -G_2 & G_2+sC_2 & -sC_2 \\ 0 & -G_3 & -sC_2 & G_3+sC_2 \end{bmatrix} \end{array},$$

and therefore,

$$Y = \begin{array}{c} \\ 1 \\ 2 \\ 3 \end{array}\begin{array}{c} \begin{array}{ccc} 1 & 2 & 4 \end{array} \\ \begin{bmatrix} G_1 & -G_1 & 0 \\ -G_1 & G_1+G_2+G_3+sC_1 & -G_3 \\ 0 & -G_2 & -sC_2 \end{bmatrix} \end{array},$$

and the transmittance from 1 to 4,

$$H_{14} = Y^{14}/Y^{11} = \frac{G_1G_2}{\begin{vmatrix} G_1+G_2+G_3+sC_1 & -G_3 \\ -G_2 & -sC_2 \end{vmatrix}} = \frac{-1}{T_1T_2s^2+T_2s(1+R_1/R_2+R_1/R_3)+R_1/R_3},$$

where $T_\alpha = R_\alpha C_\alpha$ and $R_\alpha = 1/G_\alpha$.

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A Self-Balancing Current Meter and Recorder*

An undesirable voltage drop is usually introduced across the terminals for current measurement by conventional[1] devices. This may be caused by current flow through a meter resistance or intentionally generated across a standard resistance when and where it is desired to convert the measurement to one of voltage. This extraneous potential often causes difficulties in the measurement of current-voltage relations in the fractional volt range.

With negative feedback techniques it is possible to measure currents without introducing a voltage drop across the measurement terminals. A conventional operational amplifier may be used as a fast instrument giving an output proportional to input current while keeping the potential difference across the input terminals very close to zero by feedback.[2-5] For steady or slowly changing currents, a servo self-balancing and recording potentiometer may be converted, by minor circuit changes described in this note, into a zero input impedance current recorder.

A servo system which gives an output motion proportional to the input current, while holding the potential across its input terminals at a minimum, is shown schematically in the upper part of Fig. 1. A network composed of a standard resistor R_1 in series with a fraction of a much lower resistance slidewire R_2 is connected across the input terminals. An amplifier senses the dc potential difference between the input terminals and applies torque to the motor when this voltage differs from zero. The wiper of the R_2 slidewire will then move so that the voltage across its upper portion will balance (be equal and of opposite polarity to) the potential developed across R_1 by an input current, I_1. If I_2 is much greater than I_1, the motor and wiper movement will be proportional to I_1. R_1 determines the current range for a nominal full scale potential across R_2; R_3 provides the usual calibration adjustment.

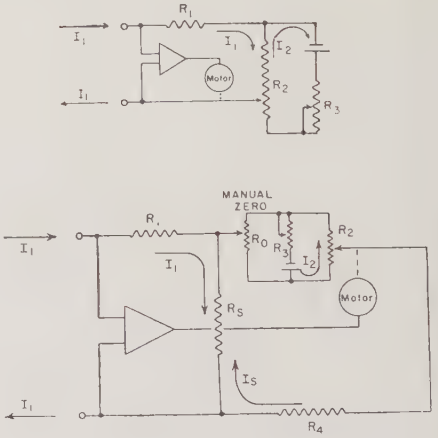

Fig. 1.

The circuit may be modified to provide for a variable zero position as well as for current ranges where a large ratio of I_2 to I_1 is inconvenient. In the configuration shown in the lower part of Fig. 1, I_1 (R_1+R_s) is made equal to R_sI_s, where I_s is proportional to the difference in the position of R_2 and the zero adjust slide contacts. The latter condition is fulfilled if R_4 is large compared to R_2, R_s,

* Received by the IRE, June 27, 1961.

[1] The work described here was completed before the recently introduced commercial devices which "clip on" rather than being inserted into the circuit.

[1] A. Nathan, "Matrix analysis of constrained networks," IEE Monograph No. 399E; September, 1960. Republished in Pt. C of *Proc. IEE*, vol. 108, pp. 98–106; March, 1961.

[2] I. Pelchowitch and J. J. Zaalberg van Zelst, "A wide-band electrometer amplifier," *Rev. Sci. Instr.*, vol. 23, pp. 73–75; February, 1952.

[3] J. Praglin, "A new high stability micromicroammeter," IRE Trans. on Instrumentation, vol. I-6, pp. 144–147; June, 1957.

[4] J. W. Moore, "Electronic control of some active bioelectric membranes," Proc. IRE, vol. 47, pp. 1869–1880; November, 1959.

[5] J. Praglin and W. A. Nichols, "High-speed electrometers for rocket and satellite experiments," Proc. IRE, vol. 48, pp. 771–779; April, 1960.

and R_0. Several commercially available self-balancing potentiometers may be converted to this "currentometer" circuit with only a few wiring changes. For preliminary tests, a simple inexpensive commercial potentiometer (Varian Assoc., Model G10, 10 mv full scale) was chosen for modification and testing. Table I gives the resistance values for full scale currents ranging from 10 μa to 0.01 μa (supplied from a source resistance of 10^6 ohms).

TABLE I

Full Scale Current	R_1	$R_4=49$ kΩ
10 μa	27 Ω	$R_3=1$ kΩ
1.0	9 kΩ	$R_0=5$ kΩ
0.1	100 kΩ	$R_2=5$ kΩ
0.01	1.0 megΩ	$R_3=1.5$ kΩ

The value of I_s varied from zero to 10 μa for full scale movement. The damping adjustment circuit normally used in the potentiometric configuration of this recorder was not required but the 1 mfd capacitor normally across R_4 was retained. The potential drop across the current measurement terminals, or "error voltage" was monitored and found to have an offset of about 200 μv and a dead-space of ± 25 to 50 μv. This dead-space amounted to about $\pm 3\times 10^{-10}$ a and was small enough to give excellent tracking and reproducibility on all except the most sensitive current range (0.01 μa full scale).

It is also possible to change the current range by setting $R_1=0$ and varying R_4. In this case, I_s is made the equal and opposite of I_1. With this circuit arrangement, the response of the modified Varian G-10 became rather sluggish in the 0.1 μa full scale range. This may have occurred because the source impedance of the balancing current ($R_4=5$ megohms) was large with respect to the input impedance of this amplifier (about 0.5 megohm). This might still be a useful circuit arrangement for use with amplifiers of higher gain and input impedance.

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Reciprocal Relations in an N-Slab Dielectric*

Transmission and reflection properties of radomes and plasma sheaths about missiles are typical of many current problems involving the propagation of radio energy through a dielectric. In such problems it is often convenient to represent the actual dielectric by a number of parallel contiguous slabs whose electrical properties are homogeneous and isotropic in each slab.

During a study in this area it became important to determine whether or not such a dielectric model exhibits reciprocal properties for its transmission and reflection coefficients. This note presents a proof that the transmission coefficients are identical for both directions of transmission but that the reflection coefficients are not.

The method and notation used follows that of Stratton.[1] He derives expressions for the transmission and reflection properties of a single slab separating two semi-infinite media having different electrical constants.

Consider a plane electromagnetic wave normally incident upon a dielectric made of n homogeneous parallel slabs, the ith slab having thickness d_i and propagation constant k_i (Fig. 1). The quantities E_I, E_R, and

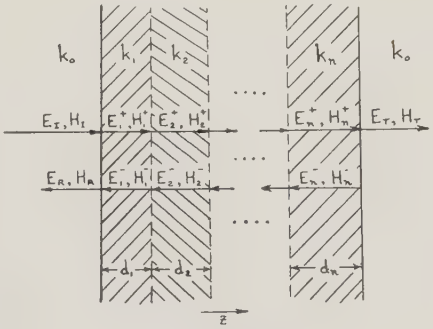

Fig. 1—Dielectric model.

E_T designate the electric intensities of the incident, reflected, and transmitted waves at their respective interfaces. The corresponding magnetic intensities are

$$H_I = \frac{k_0}{\omega\mu_0} E_I; \quad H_R = \frac{k_0}{\omega\mu_0} E_R; \quad H_T = \frac{k_0}{\omega\mu_0} E_T.$$

The total E- and H-waves in the ith slab are the sums of the forward- and backward-traveling components in this slab and are

$$\left.\begin{aligned} E_i &= E_i^+ e^{jk_i z} + E_i^- e^{-jk_i z} \\ H_i &= \frac{k_i}{\omega\mu_i}\left(E_i^+ e^{jk_i z} - E_i^- e^{-jk_i z}\right) \end{aligned}\right\} 0 \le z < d_i.$$

The expressions for the continuity of the E- and H-vectors across the ith interface (counting from left to right) are

$$e^{jk_{i-1}d_{i-1}}E_{i-1}^+ + e^{-jk_{i-1}d_{i-1}}E_{i-1}^- - E_i^+ - E_i^- = 0$$

$$k_{i-1}e^{jk_{i-1}d_{i-1}}E_{i-1}^+ - k_{i-1}e^{-jk_{i-1}d_{i-1}}E_{i-1}^- - k_iE_i^+ + k_iE_i^- = 0 \quad i = 1, 2 \cdots n+1$$

since $\mu_0=\mu_1=\mu_2=\cdots=\mu_n$. In the last pair the terms E_T, k_0E_T replace the zeros in the right members. The quantity d_0 is taken as zero. The determinant of this set of $2n+2$ linear equations is designated D. The determinants obtained by replacing the first and second columns of D by the column matrix $(0, 0, 0, \cdots 1, k_0)_{col}$ are designated N_T and N_R, respectively. The transmission and reflection coefficients T and R are defined by

$$T = \frac{E_T}{E_I}, \qquad R = \frac{E_R}{E_I},$$

so that

$$T = \frac{D}{N_T}, \qquad R = \frac{N_R}{N_T}.$$

Assume now that the wave propagates from right to left. All the E- and H-vectors, transmission and reflection coefficients, and determinants for this case are designated by primes on corresponding quantities for the opposite direction.[2] At the ith interface

$$e^{jk_id_i}E_i^{+\prime} + e^{-jk_id_i}E_i^{-\prime} - E_{i-1}^{+\prime} - E_{i-1}^{-\prime} = 0$$

$$k_ie^{jk_id_i}E_i^{+\prime} - k_ie^{-jk_jd_i}E_i^{-\prime} - k_{i-1}E_{i-1}^{+\prime} + k_{i-1}E_{i-1}^{-\prime} = 0$$

where i runs from $n+1$ to 1. The first and last pairs ($i=n+1$, $i=1$) are

$$E_I' + E_R' - E_n^{+\prime} - E_n^{-\prime} = 0$$

$$k_0E_I' - k_0E_R' - k_nE_n^{+\prime} + k_nE_n^{-\prime} = 0$$

and

$$e^{jk_1d_1}E_1^{+\prime} + e^{-jk_1d_1}E_1^{-\prime} = E_T'$$

$$k_1e^{jk_1d_1}E_1^{+\prime} - k_1e^{-jk_1d_1}E_1^{-\prime} = k_0E_T'.$$

The determinants, D, D' are equal to $(-2)^{n+1}k_0k_1k_2\cdots k_{n-1}k_n$.

We now demonstrate that $N_T=N_T'$ and that $N_R\neq N_R$. Apply the following transformations to N_T. Multiply the third, fourth, $\cdots$ $(2n+1)$th, $(2n+2)$th columns by $e^{-jk_1d_1}$, $e^{jk_1d_1}$, $\cdots$, $e^{-jk_nd_n}$, $e^{jk_nd_n}$, respectively. Interchange the first and second columns, third and $(2n+2)$th columns, fourth and $(2n+1)$th columns, etc. for a total of $(n+1)$ interchanges. Next, apply a row interchange as follows. If n is even, there will be n interchanges: first and $(2n+1)$ rows, second and $(2n+2)$th rows, third and $(2n-1)$th rows, fourth and $2n$th rows, etc. with the $(n+1)$th and $(n+2)$th rows fixed. If n is odd there will be $n+1$ interchanges of rows, the same as before, now with the $(n+1)$th and $(n+2)$th row being interchanged. Multiply all odd rows by (-1). After multiplying both the first and second columns by (-1) we obtain a determinant whose elements are identical to N_T'. Since there is always an even number of sign interchanges, we have $N_T=N_T'$ so that $T=T'$.

If these same transformations (with the exception of the last one) are applied to N_R, we obtain a determinant which differs from N_R' only in the sign of a_{11}, $a_{2n+1,2}$ elements which are now (-1). Thus, N_R' cannot equal N_R and so $R\neq R'$. This difference in the reflection coefficient is compensated by the change in the absorption of the dielectric slabs.

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* Received by the IRE, July 10, 1961.

[1] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 511–516; 1941.

[2] The wave is now incident upon the $(n+1)$th interface.

Construction of Probability Densities from Their Moments*

Most engineering problems involving stochastic variables lead to Gaussian distributions or distributions derived by functional mappings of Gaussian distributions. It is easy to show how dreadful it would be if such were not the case. The *a priori* simple problem of the construction of a probability density from the knowledge of its moments is a case in point. This is not a trivial problem because the results of a number of measurements can only lead to two interpretations: 1) an estimate for the probability of occurrence of a given value provided this value did occur and 2) an estimate of the moments. The second interpretation is less restrictive and is less subject to variations when a new set of measurements becomes available. The accuracy of the estimates of the moments, however, is poorer the higher the degree of the moment, so that in practice only a finite number of moments can be estimated.

With this in mind, let us try to find the probability density $p(r)$ when the moments m_i are given from m_1 to m_n.

It is well known that the knowledge of a number of moments of a probability density does not lead, in general, to more accurate prescriptions for the construction of this density than the Dieudonné-Markov inequalities. In general, one cannot form a sequence of approximations,

$$p_n(r) = f_n(m_1, m_2 \cdots m_n, r), \tag{1}$$

converging toward the density $p(r)$ when n increases indefinitely. Nor can one, in general, form a convergent series of the form

$$p(r) = \sum_{n=1}^{\infty} p_n(r). \tag{2}$$

Any probability density, however, has a Fourier transform,

$$\bar{p}(f) = \int_{-\infty}^{+\infty} p(r) \exp(-2\pi i f r) dr. \tag{3}$$

The exponential can be expanded in power series of the argument. Integrating, one obtains

$$\bar{p}(f) = \sum_n (-2\pi i)^n \frac{f^n m_n}{n!}. \tag{4}$$

The series (4) is convergent provided the moments do not increase too rapidly with n.

The convergence will depend on the behavior of $p(r)$ for large magnitudes of r. Even as rapidly decreasing a function as the Gaussian distribution does not yield a convergent series. On the contrary, suppose that $p(r)$ has a compact domain such that $p(r)$ is certainly zero for

$$|r| > R. \tag{5}$$

One can write

$$r = R\theta \tag{6}$$

$$p(r) = p(\theta) \tag{7}$$

$$m_n = R^{n+1}\mu_n \tag{8}$$

$$\mu_n = \int_{-1}^{+1} p(\theta)\theta^n d\theta. \tag{9}$$

* Received by the IRE, July 12, 1961.

Using vertical bars for absolute value, one obtains

$$|\mu_n| \leq \int_{-1}^{+1} p(\theta)|\theta|^n d\theta. \tag{10}$$

The upper bound [right-hand side of (10)] for the absolute value of μ_n tends toward zero when n increases indefinitely. The corresponding series (4) is convergent.

Even when the series (4) is convergent, its inverse Fourier transform P is paradoxical;

$$P(r) = \sum_n \frac{(-1)^n}{n!} \delta_0^N(r) m_n \tag{11}$$

where δ_0^N is the nth derivative of the Dirac delta function placed at the origin. The function $P(r)$ has the same moments as $p(r)$ but is not equal to $p(r)$. This is possible because $P(r)$ is not a probability density and is not positive definite.

In order to transform $P(r)$ into a probability density, one can use the Taylor expansion of the Dirac delta function placed at a distance r_i from the origin;

$$\delta_{r_i}(r) = \sum_n \frac{(-1)^n}{n!} r_i^n \delta_0^N(r). \tag{12}$$

The Taylor expansion is sufficiently correct in distribution theory,[1] because one can write

$$\delta_{r_i}(\phi) = \phi(r_i) = \sum_n \frac{r_i^n}{n!} \phi^N(0). \tag{13}$$

The functions ϕ are the usual functions of the class D. The Taylor series of a ϕ function is well defined nearly everywhere. The series is, however, not defined on the boundary of the compact domain of the function. Thus, (13) is correct for almost all ϕ functions, but not for all and this is why the word sufficiently is added. We now write

$$P(r) = \sum_i a_i \delta_{r_i}(r), \tag{14}$$

where the a_i are either positive constants or zero. Using the Taylor expansion (12)

$$P(r) = \sum_n \frac{(-1)^n}{n!} \left(\sum_i a_i r_i^n \right) \delta_0^N(r). \tag{15}$$

Equating (11) and (15),

$$\sum_i a_i r_i^n = m_n. \tag{16}$$

The function $P(r)$ (14) has the same moments as $p(r)$ provided the set of conditions (16) is satisfied. If one adds to this set of conditions that all a_i must be positive, the function $P(r)$ becomes a possible probability density. Because two probability densities which have the same moments are the same, one must have

$$a_i = \epsilon p(r_i). \tag{17}$$

The parameter ϵ is an infinitely small quantity such that $\epsilon\delta$ is the Kronecker delta. The introduction of ϵ in distribution theory permits one to define unambiguously the product of distributions and thus to generalize this essentially linear theory. In practice one does not know the higher moments with any accuracy so that one is forced to construct $p(r)$ from a finite number of moments. The set of conditions (16) is a linear inhomogeneous set in a_i. The characteristic determinant is a Van der Monde determinant. It can only be zero if two r_i are chosen equal. Thus, the set of conditions (16), restricted to a finite number of a_i equal to the finite number of available moments, has always a well-defined solution for any chosen set of r_i. The restriction that the a_i must all be positive, restricts the possible choice of r_i. Because of the essential additivity of moments, sums of properly weighted solutions are also solutions. Thus assuming all solutions obtained to be equally probable, one finds for each value of r a probability density $p_r^*(p)$ for the value of the probability density $p(r)$.

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[1] L. Schwartz, "Theorie des Distributions," Hermann and Co., Paris, France, vols. 1 and 2; 1950.

A Simple Passive-Element Electrical Analog to a Gyro*

The writer has found a very simple electrical analog to a free gyroscope which, to his knowledge, has not been previously reported in the literature.

Consider a gyroscope system with principal axes in the directions x, y, and z. Let the angular momentum of the gyro wheel be z directed and of magnitude Ω, the principal moments of inertia be I_x and I_y (including gimbals), the angular velocities be ω_x and ω_y, and external torques be T_x and T_y. The linearized equations of motion are:

$$I_x\dot{\omega}_x + \Omega\omega_y = T_x$$

$$I_y\dot{\omega}_y - \Omega\omega_x = T_y.$$

In matrix form:

$$\begin{vmatrix} \frac{I_x}{\Omega} s & 1 \\ -1 & \frac{I_y}{\Omega} s \end{vmatrix} \begin{vmatrix} \omega_x \\ \omega_y \end{vmatrix} = \begin{vmatrix} \frac{T_x}{\Omega} \\ \frac{T_y}{\Omega} \end{vmatrix}.$$

The above form clearly shows[1] why a passive element electrical analog in which both gyro equations are the analogs of loop or node equations cannot be found. The off-diagonal terms of the matrix would represent mutual impedances or mutual admittances, and hence would have to be identical. The matrix can be reordered, bringing the terms 1, -1, onto the diagonal, but a passive-element analog still cannot be found, since the -1 represents a negative self-impedance or admittance, not realizable with passive elements.

* Received by the IRE, July 17, 1961.

[1] J. A. Aseltine, "Transform Methods in Linear Systems Analysis," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 75–82; 1958.

Consider the circuit shown in Fig. 1. The node equation is

$$i + C\frac{dV}{dt} = i_s.$$

The loop equation for the L-C loop is

$$L\frac{di}{dt} - V = V_s.$$

In matrix form

$$\begin{vmatrix} CS & 1 \\ -1 & LS \end{vmatrix} \begin{vmatrix} V \\ i \end{vmatrix} = \begin{vmatrix} i_s \\ V_s \end{vmatrix}.$$

Comparison shows that the following analogue is valid,

$$C = \frac{I_x}{\Omega} \qquad L = \frac{T_y}{\Omega}$$

$$V = \omega_x \qquad i = \omega_y$$

$$i_s = \frac{T_x}{\Omega} \qquad V_s = \frac{T_y}{\Omega}.$$

For gyro systems of ordinary dimensions, realizable values of L and C would result.

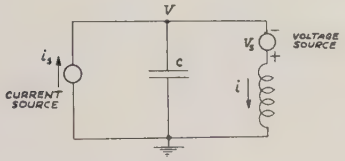

Fig. 1.

A floated gyro, floated around the x axis with viscous constant D, can be represented by addition of a resistor, $R=\Omega/D$, in parallel with C.

A rate gyro, restrained around the x axis with a spring of constant K, can be represented by adding an inductor, $L=\Omega/K$, in parallel with C.

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Measurement of Tunnel-Diode Conductance Parameters*

An accurate knowledge of the instantaneous incremental conductance (g_d) of a tunnel diode that is operating as a mixer (down converter) is necessary in order to correlate the theoretical and experimental data of such a diode's performance. This note briefly describes a method of determining g_d that has proven useful in work on tunnel-diode mixers at Philco's Western Development Laboratories.

This method assumes that the local oscillator voltage v_1 is sinusoidal and that

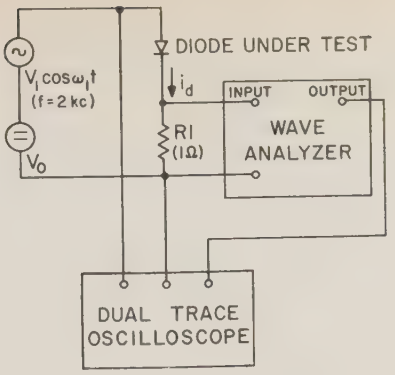

Fig. 1—Principle of test setup.

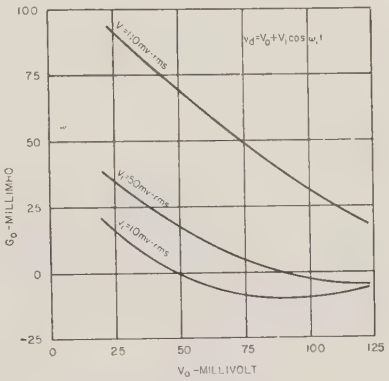

Fig. 2—G_0 vs diode bias.

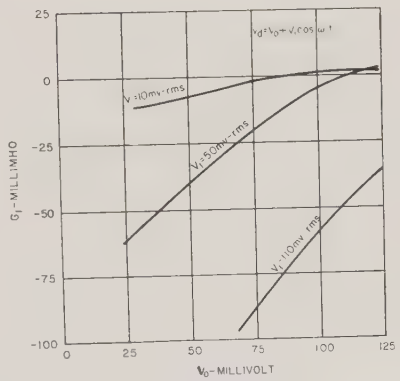

Fig. 3—G_1 vs diode bias.

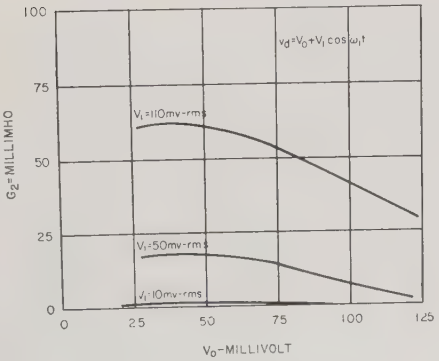

Fig. 4—G_2 vs diode bias.

g_d is a function of v_1 for a given direct voltage bias, V_0. We may therefore write

$$v_1 = V_1 \cos \omega_1 t$$

and

$$g_d = G_0 + G_1 \cos \omega_1 t + \cdots + G_n \cos n\omega_1 t,$$

where the coefficients $G_0, G_1, \cdots, G_n$ depend on only V_0 and V_1 for a given tunnel diode. The determination of g_d is based on the relationship between g_d and the current, i_d, through the diode. This relationship is given by

$$G_0 = \frac{1}{V_1}\left[I_1 + 3I_3 + 5I_5 + \cdots + \begin{cases} nI_n & (n \text{ odd}) \\ (n+1)I_{n+1} & (n \text{ even}) \end{cases} \right],$$

$$G_1 = \frac{2}{V_1}\left[2I_2 + 4I_4 + 6I_6 + \cdots + \begin{cases} (n+1)I_{n+1} & (n \text{ odd}) \\ nI_n & (n \text{ even}) \end{cases} \right],$$

$$G_2 = \frac{2}{V_1}\left[3I_3 + 5I_5 + 7I_7 + \cdots + \begin{cases} nI_n & (n \text{ odd}) \\ (n+1)I_{n+1} & (n \text{ even}) \end{cases} \right],$$

$$\vdots$$

$$G_n = \frac{2}{V_1}(n+1)I_{n+1},$$

where $I_1, I_2, \cdots, I_{n+1}$ are the coefficients of the Fourier series representation of i_d, *i.e.*,

$$i_d = I_0 + I_1 \cos \omega_1 t + \cdots + I_{n+1} \cos (n+1)\omega_1 t.$$

Thus g_d may be determined by a measurement of V_1 and of the coefficients I_q, where $q=1, 2, \cdots, n+1$.

The test setup used for the measurement of I_q is shown, in principle, in Fig. 1. The dual-trace oscilloscope is used to determine the sign of I_q, while the magnitude of I_q is determined by dividing R_1 into the voltage reading on the wave analyzer. The value of R_1 is chosen so small that the voltage drop across it has a negligible effect on the accuracy of the measurements. Some of the results obtained with this setup for a GE type ZJ56 diode are shown in Figs. 2–4.

Some values of G_0 and G_1 that were measured in this manner at audio frequencies have been found to be applicable at UHF frequencies. Thus, good agreement was found between the measured and calculated conversion gain and noise figure for a tunnel diode operating at 420 Mc. Further, this agreement was obtained not only at a single point but over a range of bias and local oscillator signal levels.[1]

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* Received by the IRE, July 5, 1961. Performed under Contract AF 04(647)-532, AFBMD, Los Angeles, Calif.

[1] For more details, see "Semiannual Report of Advanced Equipment Studies," Philco Corp., Palo Alto, Calif., WDL-TR-1481, Section I; February 15, 1961.

Experiments with Nonreciprocal Parametric Devices*

It has been pointed out by Kamal[1] that parametric diodes may be used to realize nonreciprocal networks. Recently, Baldwin[2] proposed a unilateral amplifier employing two parametric diodes pumped in quadrature.

For some time now we have been experimenting along similar lines with various nonreciprocal parametric devices. The following is a report on the results of this work.

The first of these experimental devices (Fig. 1) is a 200-Mc unilateral amplifier of the sort described by Baldwin. The two signal circuits consist of strip line resonators A and A' coupled by a quarter wavelength section of a 50-ohm cable. The common idler circuit also consists of a strip transmission line BB' capacitively loaded at both ends by parametric diodes C and C' which are also part of the signal circuits. Signal and idler circuits are separated by sections of coaxial line D and D' acting as short circuits for idler and pump frequencies and as additional capacitive loads at the signal frequency.

Pump power at 850 Mc is supplied to the diodes through small capacitors E and E' which are tuned out by coaxial stubs F and F'. A 3-db directional coupler provides the necessary 90° phase shift in the pump input circuits. At the pump frequency these circuits are separated by means of a tunable stub G in the center of the idler circuit; dc bias is also supplied at this point.

The idler circuit may be tuned from 600 Mc to 700 Mc by varying the bias voltage.

We used a matched pair of Microwave Associates' diodes No. MA-4202X. Although we were able to achieve satisfactory nonreciprocal amplification, it was observed that for a given amount of pump power, maximum gain occurred at the edge rather than in the center of the signal band.

Shifting the response back to the center of the band by detuning the idler circuit, more pump power was required for the same gain. Detailed analysis reveals the existence of a particular combination of nonzero idler and signal circuit reactances which requires less pump power than the case in which all reactances are tuned out. In this detuned condition, however, the device is no longer matched and the gain in the backward direction is larger than unity. Consequently, we might expect the noise figure to be worse than in the ideal condition.

In our case we measured a 4-db noise figure; the predicted noise figure was of the order of 1–2 db.

In agreement with the theory, we found the device to have a constant gain-bandwidth product.

The second device with which we have experimented is shown in Fig. 2. This is a 100-kc circulator. The double-tuned circuit formed by L_1 and L'_1, together with the parametric diodes D and D' is tuned to 2.5

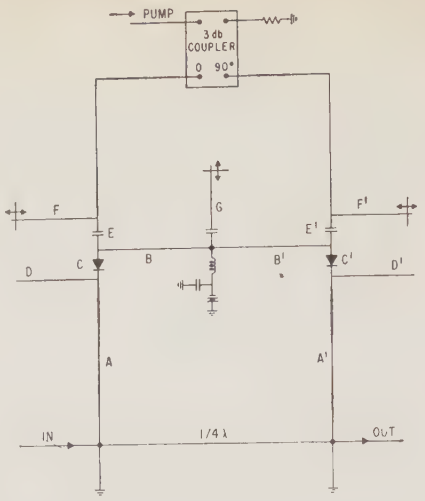

Fig. 1—Circuit diagram of 200-Mc unilateral amplifier.

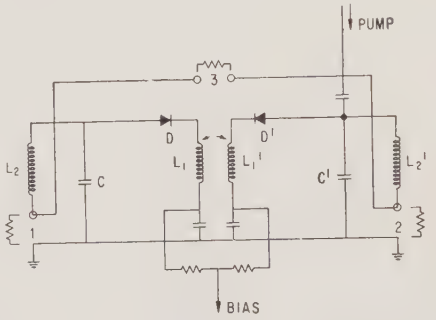

Fig. 2—Circuit diagram of 100-kc circulator.

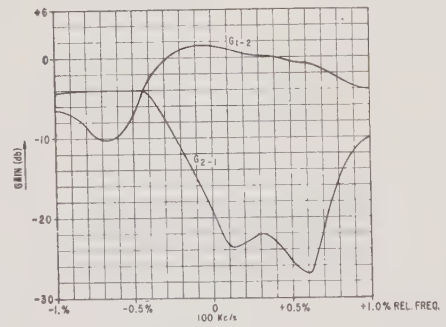

Fig. 3—Gain and isolation characteristics of 100-kc circulator.

Mc and has a bandwidth of approximately 200 kc. This circuit is excited by the pump at 2.5 Mc and will support both upper and lower sidebands.

It is separated from the signal circuits by means of bypass capacitors C and C' which are tuned out by 100 kc by L_2 and $L'_{2'}$.

The three ports of the circulator are indicated by 1, 2 and 3. By properly matching port 3 we were able to achieve isolation greater than 20 db over a 0.7 per cent band, while forward insertion gain varied from +1 db to −1 db (see Fig. 3).

It is a pleasure to acknowledge Dr. R. Adler's support and valuable suggestions.

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* Received by the IRE, August 7, 1961.

[1] A. K. Kamal, "A parametric device as a nonreciprocal element," Proc. IRE, vol. 48, pp. 1424–1430; August, 1960.

[2] L. D. Baldwin, "Nonreciprocal parametric amplifier circuits," Proc. IRE, vol. 49, p. 1075; June, 1961.

New Form of Plasma Dispersion Formula Usable for Analog Computing*

For electric analog computation of Sellmeier's dispersion formula[1] of a Lorentz plasma in the absence of collisions and of a magnetic field,

$$\mu^2 = 1 - X = 1 - \frac{f_N^2}{f^2} \tag{1}$$

(f=wave frequency and f_N=plasma frequency) can very easily be represented with sin- and cos-generators. This is not so when the magnetic field influence is introduced. Indeed, the Appleton[2]-Lassen[3] formula looks rather discouraging when

$$\mu^2 = 1 - X \Big/ \left\{1 - iZ - \frac{\frac{1}{2}Y_T^2}{1 - X - iZ} \pm \left[\frac{\frac{1}{4}Y_T^4}{(1 - X - iZ)^2} + Y_L^2\right]^{1/2}\right\} \tag{2}$$

($Y_T = f_H/f \cos\theta$, $Y_L = f_H/f \sin\theta$, $Z = \nu/2\pi f$, f_H gyrofrequency in the magnetic field B, ν effective collision number of electrons, θ angle between wave normal and magnetic field B). Now, as (1) is the asymptotic form of (2) for $f \gg f_H$, $Z \ll 1$, it is useful to write (2) in the form of (1) with an appropriate correction term M, *viz.*,

$$\mu^2 = 1 - X/(1 - iZ - M) \tag{3}$$

where

$$M = \frac{\frac{1}{2}Y_T^2}{(1 - X - iZ)} \mp \left[\frac{\frac{1}{4}Y_T^4}{(1 - X - iZ)^2} + Y_L^2\right]^{1/2} \tag{4}$$

When the square root is eliminated, the terms with Y_T^4 cancel out and we have for M the simple equation

$$M^2 - M\,\frac{Y_T^2}{(1 - X - iZ)} - Y_L^2 = 0 \tag{5a}$$

or

$$M^2 \cdot (1 - X - iZ) - M \cdot Y_T^2 - (1 - X - iZ)\,Y_L^2 = 0. \tag{5b}$$

The two roots $M+$ and $M-$ correspond to the two principal polarizations, $M+$ to the electronic[4] *extraordinary* one and $M-$ to the ionic[4] *ordinary* one. In the case $Z=0$ (no collisions), it is easily seen that in the X range admitted for propagation ($0 \leq X < 1$) we have

$$M_+ > |\,Y_L| \quad \text{and} \quad |\,Y_L| < M_- < 0. \tag{6}$$

* Received by the IRE, July 5, 1961. This work was supported by a research grant from Bundes Wirtschafts Ministerium, no. A 190, 1959.

[1] W. Sellmeier, see P. Drude, "Lehrbuch der Optik," 1900.

[2] E. V. Appleton, "The influence of the earth's magnetic field in wireless transmission," *Proc. Union Radio Sci. Internatl.*, 1 (1928), 2 (lecture presented October 13, 1927, Washington, D. C.).

[3] H. Lassen, "Über den Einfluss des erdmagnetischen Feldes auf die Fortpflanzung der elektrischen Wellen der drahtlosen Telegraphie in der Atmosphäre," *Elek. Nach.-Tech.*, vol. 4, pp. 324–334; August, 1927.

[4] R. W. Larenz, "Beitrag zur Magneto-Hydrodynamik kompressibler Medien," lecture presented at Bad Salzuflen, Germany; April, 1952 (unpublished).

For analog computing, we suppose $Z=0$. It is not interesting to use the analytical solution of (5). However, it is easier to represent (5b) directly in the form

$$MY_T^2 = (1 - X)M^2 - Y_L^2. \qquad (7)$$

It is seen from (6) that in (4) the expression in brackets should get the positive sign for M_+ and the negative for M_-. So as $(1-X)>0$, the distinction between the principal components can simply be made by means of a diode giving one or the other sign in (7).

M could be computed from X, Y_T, Y_L; but with our analog circuit representing (7), we actually use f^2, $(1-X)/f_T^2$, f_L^2 (Fig. 1,

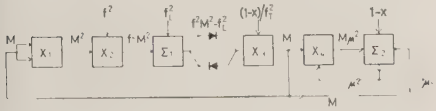

Fig. 1.

left side). The square of the refraction index μ^2 is finally calculated from (3) in the form

$$\mu^2 = 1 - X - M(1 - \mu^2) \qquad (8)$$

with the circuit of Fig. 1, right side, with the inputs: M, $(1-X)$. The total arrangement used only four multipliers and two operational amplifiers. The diode switch enables the distinction between both magneto-ionic components.

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The Expanded Reactance Series Resonator—"ERSER"*

Synopsis

ERSER is a circuit element wherein the bandwidth between the infinities of a series resonant circuit constructed from transmission line elements can be exchanged for increased steepness of the reactance slope through resonance. It finds application in B.P. filters made of T.L. elements where a compact unit of high reactance slope with change of frequency is required to be made with practical values of the characteristic impedance. In freedom from spurious responses and low insertion loss, it is superior to the alternative method of using a S.C. T.L., an integral number of $\lambda/2$ long at the resonant wavelength λ_0.

Referring to Fig. 1, ERSER consists of a S.C. transmission line of low-characteristic impedance Z_{O_1}, especially made to have high Q, terminating a T.L. of C.I. Z_O, each line being effectively $\lambda_0/4$ long. Air is used as the dielectric. Trimming adjustment can be provided by a thread T at the low-current center of the unit. The input impedance of ERSER at $100K$ per cent above the series resonant frequency, writing $\rho \equiv Z_O/Z_{O_1}$, is

$$Z_S = jZ_O \tan K\pi/2\left[\frac{1+\rho^{-1}}{\rho^{-1} - \tan^2 K\pi/2}\right].$$

This is a series resonant circuit with $Z_S=0$ at $K=0$. As the reactance of a $\lambda/2$ S.C. transmission line at $100K$ per cent above resonance is equal to $jZ_O \tan K\pi$ and is the best that can be done with a single element, the reactance scaling factor of ERSER is substantially equal to $(\rho+1)/2$; and infinities of Z_S occur when $\tan K\pi/2 = \rho^{-1/2}$ [Figs. 2(a) and 2(b)].

The step up of the current in the S.C. end of Z_{O_1} in the ratio of ρ compared with the input line would appear to condemn the device on the basis of insertion loss; however, it will be shown that it is superior in all practical cases to the only alternative, the $(\rho+1)/2$ of $\lambda/2$ S.C. transmission line.

If for convenience $(\rho+1)/2$ is considered integral and the unit of loss is that on $\lambda/4$ of T.L. of C.I. Z_O, then the loss of

$$\left[\frac{\rho+1}{2} \text{ of } \lambda/2 \text{ T.L. of C.I. } Z_O\right]: \text{ERSER}$$

$$::(1+\rho):\rho^2\left[1+\left(\frac{b}{a}\right)^{-1/\rho}\Big/\ 1 + \left(\frac{b}{a}\right)\right]+1$$

when allowance is made for the ρ:1 current step up in the outer $\lambda/4$ line (Fig. 3).

Plotting this in Fig. 3, ERSER is superior to an integral number of $\lambda/2$ S.C. T.L. up to $\rho=12$, when $b/a=20$. This is a good practical figure yielding $Z_O \doteqdot 180$ ohms and giving a reactance scaling-up of up to 6–7:1 compared with a $\lambda/2$ S.C. T.L. for $\rho \doteqdot 12$.

In conclusion, ERSER shows promise as a means of conveniently extending band-pass filters using T.L. elements into higher-impedance ranges than have previously been found convenient, and is useful for the B.P. correction of antiresonant antennas.

Note: Since preparing this note the writer has discovered the "compound line section,"[1] using two $\lambda/4$ lines in cascade. This is truly the inverse problem of the one studied here, in that Z_{O_1}/Z_O is made high in order to give a low-impedance wide-band short circuit.

The author wishes to thank the Australian Department of Civil Aviation for their support.

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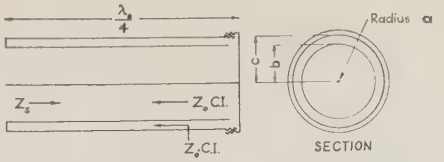
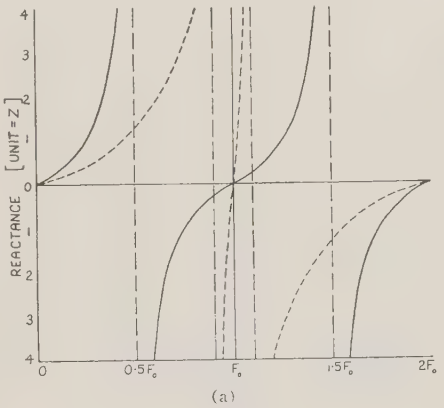
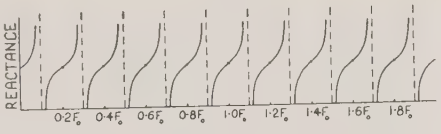


Fig. 1.—Longitudinal and cross sections of ERSER transmission line unit

$$\rho = \frac{Z_O}{Z_{O_1}}.$$



(a)



(b)

Fig. 2.—(a) Reactance curves of ERSER for $\rho=9$. Compare with 0.5 λ S.C. T.L. ($\rho=1$). Asymptotes are shown. Note bandwidth between infinities of series resonant range is reduced in ratio $2/\pi \arctan \rho^{-2}$, while reactance slope is increased in ratio $(\rho+1)/2$. (b) Reactance curves of the equivalent 2.5 λ transmission line. Note the excessive number of unwanted zeros of the reactance characteristic.

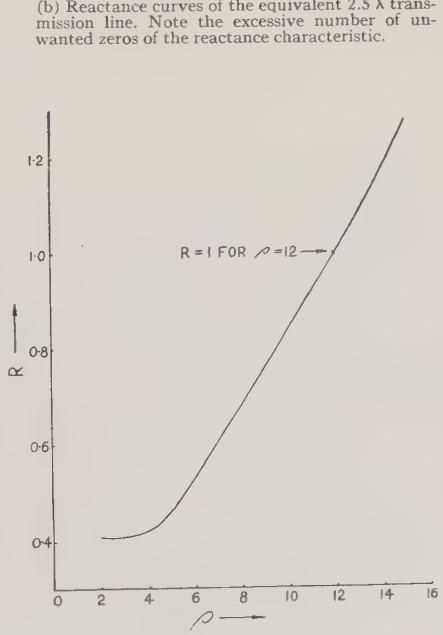


Fig. 3.—R = insertion loss ratio of ERSER compared with the 0.5 $(\rho+1)$ $\lambda/2$ long short-circuited transmission line.

* Received by the IRE, June 23, 1961.

[1] Radio Research Laboratory, Harvard University, Cambridge, Mass. "Very High Frequency Techniques," McGraw-Hill Book Co., Inc., New York, N. Y., vol. 2, p. 922; 1947.

Representation of Propagation Parameters for the Plasma in a Magnetic Field*

In a recent paper the properties of a uniform plasma for the propagation of electromagnetic waves were represented by curves plotted in the complex propagation plane.[1] The effect of a dc magnetic field can be shown by extending these curves to three-dimensional models in the way described below.

The complex propagation plane is represented by coordinates $A=(c/\omega)\alpha$ and $B=(c/\omega)\beta$, where α and β are the real and imaginary parts of the complex propagation constant and (c/ω) is the ratio of the velocity of light to the frequency of propagation. In the above-mentioned paper, the parameters plotted were the normalized collision frequency $Z=\nu/\omega$ where ν is the collision frequency and the normalized electron density $X=(\omega_p/\omega)^2$, where ω_p is the plasma frequency. The equations for the curves were derived from a conformal transformation of corresponding curves in the complex dielectric plane. These equations in turn were derived from the wave equation using the dielectric coefficient for a uniform plasma.

A later communication[2] indicated that by suitably choosing the normalized parameters to include the electron-cyclotron frequency ω_b these same curves could be used for the case of a plasma in the presence of a dc magnetic field. The purpose of this note is to point out that the original curves plotted in the complex propagation plane can be extended to three-dimensional models for the case of a dc magnetic field by introducing a third coordinate $Y=\omega_b/\omega$.

Appleton's equation for the complex refractive index of a magneto-ionic medium is given by[3]

$$n^2 = 1 - \frac{X}{1 - jZ - 1/2\dfrac{Y_y^2}{1 - X - jZ} \pm \left[1/4\dfrac{Y_y^4}{(1 - X - jZ)^2} + Y_z^2\right]^{1/2}}, \quad (1)$$

where X and Z are the normalized parameters defined above, and where

$Y_z=\omega_z/\omega$ = the normalized cyclotron frequency in the direction of propagation, z

$Y_y=\omega_y/\omega$ = the normalized cyclotron frequency in the y direction

$\omega_z=(e/m)B\cos\theta$

$\omega_y=(e/m)B\sin\theta$

θ = angle between the magnetic field B and the direction of propagation z.

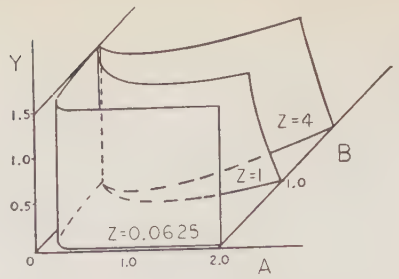


Fig. 1—Collision frequency model ordinary wave.
$A=(c/\omega)\alpha$ = normalized attenuation constant.
$B=(c/\omega)\beta$ = normalized phase-shi t constant.
$Y=\omega_z/\omega$ = normalized cyclotron f equency.
$Z=\nu/\omega$ = normalized collision frequency.

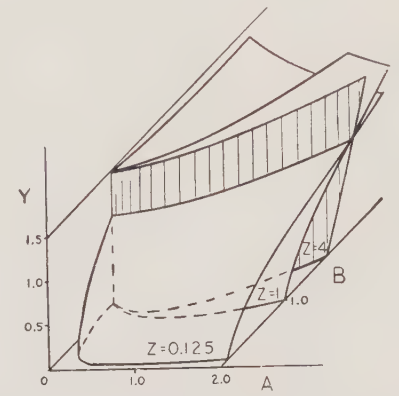


Fig. 2—Collision frequency model extraordinary wave.
$A=(c/\omega)\alpha$ = normalized attenuation constant.
$B=(c/\omega)\beta$ = normalized phase-shift constant.
$Y=\omega_z/\omega$ = normalized cyclotron frequency.
$Z=\nu/\omega$ = normalized collision frequency.

For the case of a longitud nal magnetic field, $Y_y=0$, $Y_z=Y$ and (1) reduces to

$$n^2 = 1 - \frac{X}{1 - jZ \pm Y}. \quad (2)$$

Since the square of the refractive index is equal to the dielectric coefficient, one can rationalize the right-hand side of (2) and equate the real and imaginary parts to K_r and K_i, the real and imaginary parts of the dielectric coefficient K. A conformal transformation from the complex dielectric plane to the complex propagation plane yields the following two equations for the normalized collision frequency case and for the normalized electron density case.

$$B^2 - A^2 + \frac{2AB(1 \pm Y)}{Z} = 1. \quad (3)$$

$$(A^2 + B^2)^2 - \left(2 - \frac{X}{1 \pm Y}\right)(B^2 - A^2) + \left(1 - \frac{X}{1 \pm Y}\right) = 0. \quad (4)$$

For the special case of no magnetic field these equations reduce to equations (12b) and (13b) of Bachynski.[1] They are also consistent with the redefined parameters of Bachynski.[2] Plots of (3) and (4) are shown in Figs. 1–4 where surfaces are formed for different values of Z and X. The plus sign in (3) and (4) corresponds to propagation of the ordinary wave, while the minus sign corresponds to propagation of the extraordinary wave.

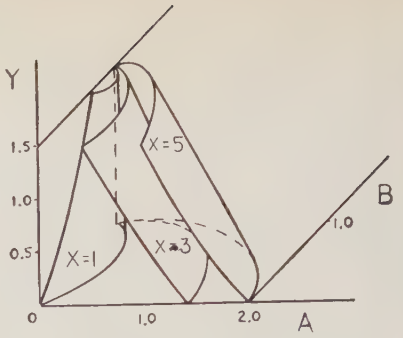


Fig. 3—Electron density model ordinary wave.
$A=(c/\omega)\alpha$ = normalized attenuation constant.
$B=(c/\omega)\beta$ = normalized phase-shift constant.
$Y=\omega_z/\omega$ = normalized cyclotron frequency.
$X=(\omega_p/\omega)^2$ = normalized electron density.

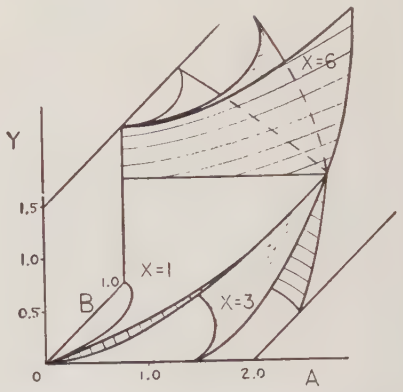


Fig. 4—Electron density model extraordinary wave.
$A=(c/w)\alpha$ = normalized attenuation constant.
$B=(c/w)\beta$ = normalized phase-shift constant.
$Y=\omega_z/\omega$ = normalized cyclotron frequency.
$X=(\omega_p/\omega)^2$ = normalized electron density.

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* Received by the IRE, June 30, 1961. This work was partially supported by NASA under research grant no. NsG 48-60.

[1] M. P. Bachynski, *et al.*, "Electromagnetic properties of high-temperature air," Proc. IRE, vol. 48, pp. 347–356; March, 1960.

[2] M. P. Bachynski, *et al.*, "Universal representation of electromagnetic parameters in presence of dc magnetic fields," Proc. IRE, vol. 49, pp. 354–355; January, 1961.

[3] J. A. Ratcliffe, "The Magneto-Ionic Theory and its Applications to the Ionosphere," Cambridge University Press, Cambridge, England; 1959.

Study of Electromagnetic Wave Polarization in Magneto-Plasmas by a Matrix Method of Crystal Optics*

In a series of papers Jones[1] has developed a method for treating the behavior of opti-

* Received by the IRE, June 30, 1961. This work was partially supported by NASA under research contract no. NsG 48-60.

[1] R. G. Jones, "A new calculus for the treatment of optical systems," *J. Opt. Soc. Am.*, vol. 31, pp. 488–503, July, 1941; vol. 32, pp. 486–493, August, 1942; vol. 37, pp. 107–112, February, 1947; vol. 38, pp. 671–684, August, 1948; vol. 46, pp. 126–131, February, 1956.

cal systems. We propose here a method for treating the study of electromagnetic wave propagation through magneto-plasmas by applying a modification of his method which is appropriate to the microwave region of the spectrum.

The method of Jones is based upon the assumption that the components of the electric field vector of the light wave leaving the system are linear functions of the electric field components of the wave entering the system. That is, if E_1 and E_2 are the complex components of a wave entering the system, then the output components of the wave, E_1' and E_2', are given by

$$\begin{bmatrix} E_1' \\ E_2' \end{bmatrix} = \begin{bmatrix} m_{11} & m_{12} \\ m_{21} & m_{22} \end{bmatrix} \begin{bmatrix} E_1 \\ E_2 \end{bmatrix}$$

or

$$E' = ME,$$

where M is a matrix characterizing the system.

Systems which act to alter the relationship between orthogonal components of an electromagnetic wave in the microwave region might usefully be characterized by such a matrix. A plasma in the presence of a dc magnetic field might be considered as such a system.

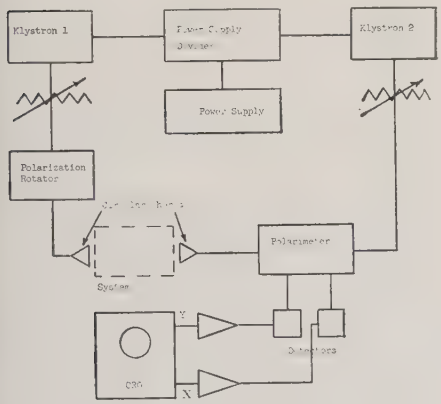

Fig. 1—Polarimeter microwave circuit.

A method for measuring the matrix has been proposed[2] for application in the optical regions. The ease with which the components of the electric field can be separated in a microwave circuit makes the measurement of the matrix M simpler at microwave frequencies.

By factoring out m_{22} three polarization measurements using a microwave polarimeter[3] will determine the matrix M to within the complex constant m_{22}. These measurements consist of measuring the output polarization for three differently oriented linearly polarized input signals. A block diagram of the microwave circuit is shown in Fig. 1. The complex constant m_{22} can be determined from a conventional attenuation and phase-shift measurement. This is done by orienting the rectangular guide of the receiving horn such that only the E_2' component is received when the input signal to the system is linearly polarized in the X_2 direction. In this way the matrix M can be uniquely determined.

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[2] *Ibid.*, see vol. 37.
[3] P. J. Allen and R. D. Tompkins, "An instantaneous microwave polarimeter," Proc. IRE, vol. 47, pp. 1231–1237; July, 1959.

A Decoding Procedure for Double-Error Correcting Bose-Ray-Chaudhuri Codes*

A two-error correcting Bose-Ray-Chaudhuri Code[1] with m information bits has as Syndromes corresponding to a single error in the $(r+1)$th bit position the vector (x^r, x^{3r}), where x is the primitive element of the $GF(2^m)$ field. A double-error would give rise to the syndrome $(x^r+x^s, x^{3r}+x^{3s})$. If we call the two components of this vector m and n respectively, we have

$$n = (x^r + x^s)(x^{2r} + x^{r+s} + x^{2s})$$

$$m^3 = (x^r + x^s)(x^{2r} + x^{2s}),$$

or

$$1 + \frac{n}{m^3} = \frac{x^{r+s}}{x^{2r} + x^{2s}} = \frac{x^{2r+(s-r)}}{x^{2r}(1 + x^{2(s-r)})}$$

$$= \frac{x^{s-r}}{1 + x^{2(s-r)}} = f(s - r) \text{ (say)}.$$

It will be noticed that

$$f(s - r) = f(2^m - 1 - s + r),$$

so that a table containing only $2^{m-1}-1$ entries would set up a one-to-one correspondence between $1+(n/m^3)$ and $s-r$.

Also, $m=x^r(1+x^{s-r})$, so that the error positions r and s can both be obtained by the given manipulation of m and n.

As an example, the double-error correcting (15, 7) Bose-Ray-Chaudhuri code has as single error syndromes,

	1000 (1)	1000 (1)
	0100 (x)	0001 (x^3)
	0010 (x^2)	0011 (x^6)
*	0001 (x^3)	0101 (x^9)
	1100 (x^4)	1111 (x^{12})
	0110 (x^5)	1000 (1)
	0011 (x^6)	0001 (x^3)
*	1101 (x^7)	0011 (x^6)
	1010 (x^8)	0101 (x^9)
	0101 (x^9)	1111 (x^{12})
	1110 (x^{10})	1000 (1)
	0111 (x^{11})	0001 (x^3)
	1111 (x^{12})	0011 (x^6)
	1011 (x^{13})	0101 (x^9)
	1001 (x^{14})	1111 (x^{12})

and the correspondence between $s-r$ and $x^{s-r}/1+x^{2(s-r)}$ is as follows:

$s-r$	$\frac{x^{s-r}}{1+x^{2(s-r)}}$
1	x^8
2	x
3	x^5
4	x^2
5	1
6	x^{10}
7	x^4

Let there be an error in the positions shown by asterisks in the syndrome matrix so that the resulting syndrome, being the sum of the corresponding rows, is

$$1100(m = x^4)0110(n = x^5),$$

yielding

$$1 + \frac{n}{m^3} = 1 + \frac{x^5}{x^{12}} = 1 + x^8 = x^2,$$

whence we have from the table of correspondence $s-r=4$, *i.e.*, there is a distance of four bits between the error positions.

Again,

$$x^r = \frac{m}{1 + x^4} = \frac{x^4}{x} = x^3,$$

that is, $r=3$ yielding the error positions as 4 and 8.

In the case of single errors, $1+(n/m^3)$ will yield 0, so that the position of the error can be found from m above.

Unlike the Peterson Decoding Method,[2] this method cannot be easily generalized to a larger number of errors. However, in the special case of double-error correcting codes, the method is advantageous in involving no successive trials.

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* Received by the IRE, July 17, 1961.
[1] R. C. Bose and D. K. Ray-Chaudhuri, "On a class of error correcting binary group codes," *Information and Control*, vol. 3, pp. 68–79; 1960.
[2] W. W. Peterson, "Error Correcting Codes," Mass. Inst. Tech. Press, Cambridge, and John Wiley and Sons, Inc., New York, N. Y.; 1961.

Nonrealizability of the Complex Transformer*

In these Proceedings Carlin[1] has given a passive (and real) realization for the complex transformer at a single real frequency, $p_0=j\omega_0$. The question arises if such a transformer can be realized at a single complex frequency p_0 with Re $p_0>0$. This question is of importance because of the recent uses of such a device in the various branches of electrical engineering. In this connection

* Received by the IRE, July 10, 1961.
[1] H. J. Carlin, "On the physical realizability of linear non-reciprocal networks," Proc. IRE, vol. 43, pp. 608–616; May, 1955. (See p. 611.)

Edelmann[2] has used them for visualizing the various transformations of power engineering. Bello[3] has used them to interpret the energy functions of nonreciprocal systems, and Belevitch[4] has incorporated them in an ingenious extension of the Brune synthesis to nonreciprocal networks. We will show that *Carlin's result does not extend to p_0 with* Re $p_0>0$.

Let a superscript asterisk denote complex conjugation and a superscript tilde denote matrix transposition. A device described by

$$V_2 = NV_1$$
$$I_1 = -\tilde{N}^* I_2, \qquad (1)$$

with the V's and I's n-vectors and N an $n\times n$ complex constant matrix, is called a complex transformer $2n$-port (here we assume that N is not real). By properly loading such a device by an n-port of admittance matrix Y_2, we can obtain an input admittance of

$$Y_1 = \tilde{N}^* Y_2 N. \qquad (2)$$

Further we have for the complex transformer itself

$$\tilde{V}_2^* I_2 + \tilde{V}_1^* I_1 = 0. \qquad (3)$$

Interpreting the components of the V's and I's as phasors, this last equation states that the average power input in the sinusoidal steady state is zero. Thus, the complex transformer has several of the useful properties of the real transformer.

We can now apply the result of Desoer and Kuh[5] which states that a network is passive at p_0 with Re $p_0\geq 0$, if and only if, $q_-(p_0)\geq 0$. For our purposes it is more convenient to use a generalization of q_- defined by

$$Q_-(V, I, p) = \begin{cases} 1/2[\tilde{V}^* I + \tilde{I}^* V] & \\ -(\sigma/|p|)|\tilde{V} I| & \text{if } \omega \neq 0 \\ 1/2[\tilde{V}^* I + \tilde{I}^* V] & \text{if } \omega = 0. \end{cases} \qquad (4)$$

Here $|\ |$ denotes the absolute value. In Newcomb,[6] it is shown that the inequality $Q_-(V, I, p_0)\geq 0$ must hold for every V and I if the network is to be passive at p_0. By using (3) and its complex conjugate transpose, we obtain for the complex transformer

$$Q_-(V, I, p) = \begin{cases} -(\sigma/|p|)|\tilde{V} I| & \text{if } \omega \neq 0 \\ 0 & \text{if } \omega = 0. \end{cases} \qquad (5)$$

Clearly, if $\sigma_0>0$, $\omega_0\neq 0$, then $Q_-(V, I, p_0)<0$ and the network is not passive at p_0. As a consequence the complex transformer can't be realized at p_0 by passive, real elements. It should be observed that $Q_-(V, I, j\omega)=0$ which, because of the synthesis method of Newcomb,[7] coincides with Carlin's result. Further for $\sigma_0>0$, but $\omega=0$, the device doesn't have real elements and can't be realized at a fixed frequency $p_0=\sigma_0$ even though $Q_-(V, I, \sigma_0)=0$. An alternate derivation of the "activity" of the complex transformer is found in Newcomb.[8]

The author is indebted to the National Science Foundation for the support of this work and Profs. Kuh and Desoer for their interest and guidance.

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[7] *Ibid.*, p. 26.
[8] *Ibid.*, p. 25.

Tunnel-Diode Super-Regenerative Parametric Motor*

In an effort to harness the stimulance of a diode into rotary motion while maintaining the contact-free operation of the parametric motor, the author has carried out measurements on tunnel diodes and built a number of electro-mechanical devices.[1-3] The basic case of resistive load and maintained stability is shown in Fig. 1(a). If the load consists of a tuned circuit of proper inductance and capacitance, and if the dissipation of the system is sufficiently reduced, supersonic or high audio-frequency oscillations can be obtained, provided that the bias voltage is made to exceed the peak-point value V_p, but not the valley-point value V_v. The circuitry with diode and load in series connection is then largely that shown in Fig. 1(b), for which the steady-state frequency to a first approximation may be written

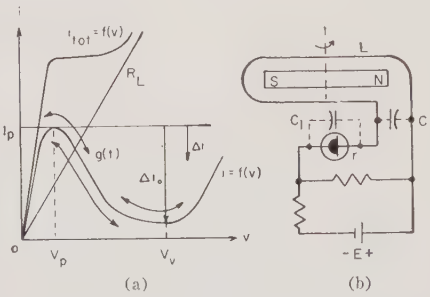

Fig. 1—(a) Tunnel-diode characteristics with possible variational bias ranges. (b) A simple scheme for turning tunnel-diode stimulance into rotary motion.

$$f_0 \doteq \frac{1}{2\pi}\sqrt{\frac{1}{LC'} - \frac{g^2}{C_1 C'}}, \qquad (1)$$

where $C'=C+C_1$. If we now make the bias voltage variational and periodic around a chosen point, such as the peak point or the valley point, the tunnel diode will never settle down to the value f_0 in (1); on the contrary, it produces a Fourier spectrum of frequencies. Swinging into the negative-slope region in Fig. 1(a), we encounter a growing transient wave train with an instantaneous amplitude associated with the downwards current deflection Δi. The wave-train instantaneous amplitude $i(t)$ is in the interval of interest almost proportional to Δi, so that we may write to a first approximation $i(t)\doteq K[1-\exp(-t/T)]$, where T is the time constant pertaining to the supersonic wave-train envelope. We observe that the generated stimulance is far from constant, taking the form of the time function

$$g(t) = g + \sum_{p=1}^{\infty} g_{p\max}\cos(p2\pi ft + \phi_p). \qquad (2)$$

The net effect of a periodic bias voltage is a varying negative conductance, going through a maximum, the subsequent quenching action that makes the magnetic field appear and vanish in a periodic fashion, and the ensuing supersonic growing and decaying wave train of variational frequency. Utilizing these facts for the accomplishment of rotary motion, we may now formulate a simple question: how can we make an alternating field in an inductor do mechanical work in such a manner that a periodic induced EMF and bias voltage results, all within the realms of the source power, secured from the negative-conductance slope of the tunnel diode? The simple answer appears in Fig. 1(b). If here the inductor contains a spinning magnet NS, the pulses of magnetic field will maintain the magnet in rotation, with the induced EMF of the spinning magnet providing the periodic bias voltage. This new electric motor has only three essential parts: the inductor L with its winding capacitance C, the rotor-magnet NS, and the tunnel diode.[4] The speed of the motor is governed by the time constant T of the wave-train envelope, and the speed is therefore quite constant. Due to the quenching action the operation is said to be superregenerative; the response on a CRO is quite typical for that of a super-regenerative device.

Parametric action appears in more than one respect. For an elementary system that may be represented by a single mesh with inductance and capacitance, loss resistance R, and slope-achieved stimulance R_N, there is a considerable change in inductance due to the rotation of the magnet, as is partly evident from the extensive change in pitch of the supersonic oscillation. A first approach integro-differential equation may therefore be written in the form

$$L\frac{di}{dt} + \left(\frac{dL}{dt} + R_n + R\right)i + \Gamma\int i\,dt = e(t). \qquad (3)$$

[4] Made by Sine-Ser Co., Waltham, Mass.

[2] H. Edelmann, "Über die Anwendung von Übertragermatrizen in untersuchungen auf dem Netzmodell," *Arch. elekt. Übertragung*, vol. 11, pp. 149–158; April, 1957.
[3] P. Bello, "Extension of Brune's energy function approach to the study of LLF networks," IRE Trans. on Circuit Theory, vol. CT-7, pp. 270–280; September, 1960. (See p. 273.)
[4] V. Belevitch, "On the Brune process for n-ports," IRE Trans. on Circuit Theory, vol. CT-7, pp. 280–296; September, 1960. (See p. 284.)
[5] C. A. Desoer and E. S. Kuh, "Bounds on natural frequencies of linear active networks," *Proc. of the Polytechnic Inst. of Brooklyn Symp.*, vol. X, pp. 415–436; 1960. (See p. 425.)
[6] R. W. Newcomb, "Synthesis of Networks Passive at p_0," University of California, Berkeley, ERL Rept., Series 60, Issue No. 317; September, 1960. (See p. 2.)

* Received by the IRE, July 20, 1961.
[1] H. E. Stockman, "Parametric oscillatory and rotary motion," Proc. IRE (Correspondence), vol. 48, pp. 1157–1158; June, 1960.
[2] H. E. Stockman, "Parametric variable-capacitor motor," Proc. IRE (Correspondence), vol. 49, pp. 970–971; May, 1961.
[3] H. E. Stockman, "Electric field motor" and "Tunnel diode electromechanical movement," U. S. Patent applications of January 26, 1961, and July 12, 1961.

We may here rely on dL/dt and R_N to cancel R, but in a different mode of operation we may rely on R_N alone to cancel the sum of dL/dt and R. The peculiar modes and the mode-switching of this motor leads to rather intricate equations in any effort of analyzing the motor operation.

This is probably the world's simplest dc motor and electronics motor, but it is essentially of interest because of its constant speed and "indestructible" design. One can actually make the speed increase slightly as the battery ages. The tunnel diode is almost independent of nuclear radiation; a fact of interest in space applications. Generally the motor then has small power requirements, and may be operated from thermoelectric generators, atomic batteries, and solar cells of the type used in space vehicles. If a restoring force is applied to the moving magnet, pendulum oscillations result. These are of theoretical interest since they start to build up from building-noise level.

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An Adaptive Communications Filter*

Considerable interest has been expressed recently in the application of adaptive filters to communication problems.[1] The literature has described techniques that allow a matched filter for an unknown repetitive signal to be synthesized using information obtained from measurements of the signal and noise. Perhaps a more useful filter might be one which is optimum for a completely random communication signal (such as a random telegraph wave). Here the problem is not "optimum detection" as in a matched filter problem, but "optimum filtering."

Such a filter is described in this note. It is one of several known techniques that attempt to adjust for minimum mean-square error. Assume that the filter is to be of the form shown in Fig. 1. It is desired to adjust the potentiometers, a_i, to give minimum mean-square error. The number of taps, k_0, is fixed. The transfer function of the filter is

$$G(Z) = \sum_0^{k_0} a_n Z^{-n} \tag{1}$$

The equation for the mean-square error may be obtained from the block diagram of Fig. 2. This equation is well known and is[2]

$$\overline{e^2} = R_S(0) + \frac{T}{2\pi j}\oint_\Gamma \cdot \left\{S_{ii}(Z)\left[\sum_0^{k_0} a_n Z^{-n}\sum_0^{k_0} a_n Z^{n}\right] - \sum_0^{k_0} a_n(Z^{-n}+Z^{n})S_{ss}(Z)\right\}\frac{dZ}{Z} \tag{2}$$

where $R_S(\tau)$ is the signal autocorrelation function, $S_{ii}(Z)$ is the sampled noise plus signal spectrum, $S_{ss}(Z)$ is the sampled signal spectrum, and the integration is around the unit circle. Uncorrelated signal and noise is assumed.

The optimum setting for a_i may be found by setting the partial of $\overline{e^2}$ with respect to a_i equal to zero and solving for a_i:

$$\frac{\partial\overline{e^2}}{\partial a_i} = \frac{T}{2\pi j}\oint_\Gamma \frac{dZ}{Z}\left\{S_{ii}(Z)\left[Z^{i}\sum_0^{k_0} a_n Z^{-n} + Z^{-i}\sum_0^{k_0} a_n Z^{n}\right] - S_{ss}(Z)\left[Z^{i}+Z^{-i}\right]\right\}$$

$$= \frac{T}{2\pi j}\oint_\Gamma \frac{dZ}{Z}\left\{S_{ii}(Z)Z^{i}\left[\sum_0^{k_0} a_n Z^{-n} - \frac{S_{ss}(Z)}{S_{ii}(Z)}\right]\right\} + \frac{T}{2\pi j}\oint_\Gamma \frac{dZ}{Z}\left\{S_{ii}(Z)Z^{-i}\left[\sum_0^{k_0} a_n Z^{n} - \frac{S_{ss}(Z)}{S_{ii}(Z)}\right]\right\} = 0. \tag{3}$$

Parseval's theorem for discrete systems is

$$\frac{1}{2\pi j}\oint_\Gamma F_1(Z)F_2(Z^{-1})\frac{dZ}{Z} = \frac{1}{2\pi j}\oint_\Gamma F_1(Z^{-1})F_2(Z)\frac{dZ}{Z} = \sum_{-\infty}^{+\infty} f_1(nT)f_2(nT). \tag{4}$$

Then a solution to (3) is

$$\frac{1}{2\pi j}\oint_\Gamma \frac{dZ}{Z}\left\{S_{ii}(Z)Z^{i}\left[\sum_0^{k_0} a_n Z^{-n} - \frac{S_{ss}(Z)}{S_{ii}(Z)}\right]\right\} = \frac{1}{2\pi j}\oint_\Gamma \frac{dZ}{Z} S_{ii}(Z)\sum_0^{k_0} a_n Z^{i-n} - \frac{1}{2\pi j}\oint_\Gamma S_{ss}(Z)Z^{i}\frac{dZ}{Z} = 0. \tag{5}$$

Again using Parseval's theorem

$$\frac{1}{2\pi j}\oint_\Gamma \frac{dZ}{Z} S_{ii}(Z)\sum_0^{k_0} a_n Z^{i-n} = \sum_{n=0}^{k_0} R_{S+N}[(n-i)T]\frac{a_n}{T}. \tag{6}$$

$R_{S+N}(T)$ is the signal-plus-noise autocorrelation function. Taking the inverse,

$$\frac{1}{2\pi j}\oint_\Gamma S_{ss}(Z)Z^{i}\frac{dZ}{Z} = \frac{R_S(iT)}{T}. \tag{7}$$

Thus, we have (k_0+1) equations of the form

$$\sum_{n=0}^{k_0} a_n R_{S+N}[(n-i)T] = R_S(iT). \tag{8}$$

Expressed in matrix form this is

$$\begin{bmatrix} R_{S+N}(0) & R_{S+N}(T) & R_{S+N}(2T) & \cdots & R_{S+N}(k_0T) \\ R_{S+N}(T) & R_{S+N}(0) & R_{S+N}(T) & & \vdots \\ R_{S+N}(2T) & R_{S+N}(T) & R_{S+N}(0) & & \\ \vdots & & & \ddots & \\ R_{S+N}(k_0T) & \cdots & & & R_{S+N}(0) \end{bmatrix}\begin{bmatrix} a_0 \\ a_1 \\ a_2 \\ \vdots \\ a_{k_0} \end{bmatrix} = \begin{bmatrix} R_S(0) \\ R_S(T) \\ R_S(2T) \\ \vdots \\ R_S(k_0T) \end{bmatrix} \tag{9}$$

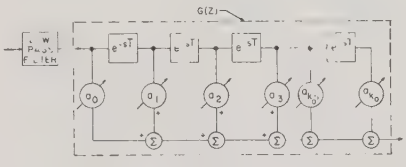

Fig. 1—A tapped delay-line filter.

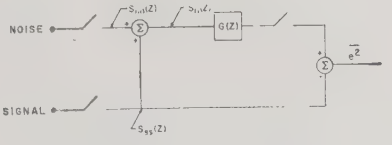

Fig. 2—The mean-squared error.

Eq. (9) shows that for a tapped delay line filter of length k_0T with no feedback, no autocorrelation for τ greater than k_0T is needed. The converse is not true, however. Even if the autocorrelations are zero for $\tau < k_1T$, the longer the length of the delay line, in general, the better the results. Eqs. (9) may be modified for some other operations on the signal. For example, if prediction is desired, $R_S(iT)$ should be replaced with $R_S[(m+i)T]$, where mT is the prediction time.

The block diagram for measuring the autocorrelation functions is shown in Fig. 3. The tapped delay line could also serve as the line used in the synthesis of the filter. Uncorrelated additive signal and noise are assumed. It is further assumed that the noise is stationary, or that its time-varying autocorrelation is known. This is a valid

* Received by the IRE, July 7, 1961.

[1] E. M. Glaser, "Signal detection by adaptive filters," IRE TRANS. ON INFORMATION THEORY, vol. IT-7, pp. 87–88; April, 1961.

[2] J. R. Ragazzini and G. F. Franklin, "Sampled-Data Control Systems," McGraw-Hill Book Co., Inc., New York, N. Y., ch. 10; 1958.

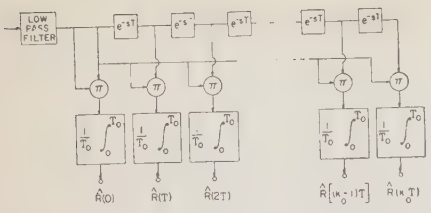

Fig. 3—Measuring the autocorrelation functions.

assumption under most practical circumstances. In fact the low-pass filter of Fig. 1 can usually be adjusted to give a noise autocorrelation that is zero for $\tau = nT$; $n \neq 0$. If the noise is stationary and no signal is present, $R_N(nT)$ is estimated and this information is stored. Then continuous estimates of $R_{S+N}(nT)$ are made. $R_S(nT)$ is

$$\hat{R}_S(nT) = \hat{R}_{S+N}(nT) - \hat{R}_N(nT). \quad (10)$$

Of course, if the signal is stationary, T_0 can be made very large and good estimates of $R_S(nT)$ will be available.

If the low-pass filter cuts off at W cps, then T, the delay between taps, should be less than $1/2W$. For a small number of taps, conceptually, the complete filter could be built on an analog computer. However, a useful filter most likely will always be constructed on a digital computer. Routines are available for easily solving 100 linear simultaneous equations, and a 100-tap delay line makes possible a complicated $G(z)$.

If $f(nT)$ is the sampled input to the computer, and $C(nT)$ is to be the sampled filter output, the $C(nT)$ and $f(nT)$ are related by the simple difference equation:

$$C(nT) = a_0 f(nT) + a_1 f[(n-1)T] + a_2 f[(n-2)T] + \cdots + a_{k_0} f[(n-k_0)T]. \quad (11)$$

This is the digital computer simulation of $G(Z)$. The sampling theorem shows

$$\frac{1}{T_0}\int_0^{T_0} f(t+nT)f(t)dt = \frac{1}{2WT_0}\sum_{m=0}^{2WT_0} f(mT)f[(m+n)T]. \quad (12)$$

Therefore, the operations of Fig. 3 can easily be simulated digitally. A low-pass filter of bandwidth W should precede the analog-to-digital converter.

In rare cases, the square matrix of (9) could be singular. For example, if a no-noise signal were periodic exactly at the sampling frequency, two rows of the matrix would be identical. These situations should be of no practical interest.

ACKNOWLEDGMENT

Special thanks are due R. Lawhorn and Dr. G. Franklin for their very helpful conversations and criticisms.

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During World War II he supervised electronic activities at the Kaiser Shipyards, Richmond, Calif. After the war he organized the Ross Radio Corporation, Los Altos Hills, Calif., which specializes in custom-made radios, television sets, and industrial electronics. In 1948 he joined the staff of the University of California Radiation Laboratory, Berkeley, where he developed electronic controls for large nuclear accelerators. He participated in a classified project for the Atomic Energy Commission overseas as a Group Leader in 1951. Upon his return he assumed charge of the Special Problems Group at the Radiation Laboratory. He is the inventor of the Kaiser-Aiken thin cathode-ray tube. He has held his present position as Director of Research for the Kaiser Aircraft and Electronics Corporation, West Coast Electronics Laboratory, Palo Alto, Calif., since 1953.

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Mr. Endresen is an Avionics Panel member of the NATO Advisory Group for Aeronautical Research and Development.

❖

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He joined the Norwegian Defence Research Establishment, Division for Telecommunication, Lilleström, Norway, in 1957 as a Research Engineer. He is active in the field of military electronics, particularly interference problems, such as the behavior of linear and nonlinear communication and radar systems subjected to heavy interference of various types.

❖

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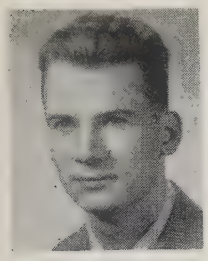
From 1951 to 1952 he worked on com-

puter circuit design at Northrop Aircraft Company, Los Angeles, Calif. In 1953 he was engaged in microwave relay research at the Bell Telephone Laboratories, Murray Hill, N. J. From 1953 to 1956 he was a Research Assistant in microwave physics at Stanford. Since 1956 he has been doing research in physical electronics at the Central Research Department of Varian Associates, Palo Alto, Calif.



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Dr. Helmer is a member of Phi Beta Kappa, Tau Beta Pi, Sigma Xi, and the American Physical Society.

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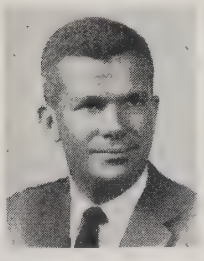

L. R. KAHN

From 1944 to 1946 he served in the U. S. Army Signal Corps, and from 1947 to 1950 he was employed by RCA Communications, Inc., and was loaned to RCA laboratories during the last year of employment. From 1950 to 1952 he was employed by Crosby Laboratories, Inc. In 1952 he organized the Kahn Research Laboratories, where he is now President and Research Director.

He has conducted research and development on a number of original communication techniques, including single-side band transmission systems, tone converters, and other communications and television equipment. He has also served as Consultant to the Voice of America, Radio Free Europe, Willys Motors, Grumman Aircraft and Electronics, and Fairchild Engine and Airplane Corporation. He has a number of inventions in the communications field, including the ratio-squarer diversity system and television and communications transmitters.

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Kurt Schlesinger (A'41–SM'51–F'54) was born in Berlin, Germany, on April 20, 1906. He received the Diplom-Ingenieur degree in 1928 and the Doctor of Engineering degree in 1929, both from the Technische Hochschule in Berlin.

K. SCHLESINGER

For the next ten years he was active in television research in Europe. From 1941 to 1944 he was with RCA Laboratories, Purdue University, Lafayette, Ind., and from 1944 through 1947 he was Consulting Engineer in color television for CBS. He was head of a TV research department at Motorola, Inc., Chicago, Ill., from 1947 to 1958. At present he is Consulting Engineer for the Cathode Ray Tube Department of the General Electric Company, Syracuse, N. Y.

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His interest in the microwave field led to a doctoral dissertation on artificial dielectrics. An article based on this dissertation, published by the *Journal of the British IRE*, earned him the Clerk-Maxwell Premium for "the most outstanding paper published in the Institution's Journal" in 1952.

During the war, he served with the 8th Air Force in Europe as an airborne-radar specialist. From 1951 to 1955 he was a Research Associate at Stanford University, Stanford, Calif., and from 1953 to 1955 he was Assistant to the Director of Stanford's Microwave Laboratory and Lecturer in the Department of Electrical Engineering. In 1955 he joined the faculty of the University of California, Berkeley, where he is now Associate Professor of electrical engineering.

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❖

Lofti A. Zadeh (S'45–A'47–SM'56–F'58) was born in Baku, Russia, on February 4, 1921. He attended the American College, Teheran, Iran, and received the B.S. degree in electrical engineering from the University of Teheran in 1942. He came to the United States in 1944 and entered the Massachusetts Institute of Technology, Cambridge, where he received the M.S. degree in 1946.

L. A. ZADEH

He joined the staff of Columbia University as an Instructor in electrical engineering and received the Ph.D. degree in 1949. He was promoted to Assistant Professor in 1950, Associate Professor in 1953, and became a full Professor in 1957. In 1956 he was a member of the Institute for Advanced Study, Princeton, N. J. At present he is Professor of electrical engineering at the University of California, Berkeley.

He has published papers on the subjects of time-varying networks, optimal filters, and nonlinear systems, with emphasis on the general principles underlying the transformation of signals and the characterization of input-output relationships. His main interest at present is in system theory.

Dr. Zadeh is Chairman of U. S. Commission VI and the International Scientific Radio Union, and a member of the AIEE, the American Mathematical Society, the American Physical Society, the Institute of Mathematical Statistics, the ACM, the Society for Industrial and Applied Mathematics, Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

Books

Wave Propagation in a Turbulent Medium, by V. I. Tatarski

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 257 pages+xiv pages+6 bibliography pages+2 appendix pages+22 pages notes and remarks. Illus. 7¼×10. $9.75.

In this extensive monograph the author not only has collected in one place the dispersed theories describing various aspects of wave propagation through irregular media; he also has attempted to relate these theories to physical processes that would likely be responsible for such irregularities of refractive index in the atmosphere. It is this tie between wave theory, on the one hand, and atmospheric physics, on the other, that will be especially valuable to research workers who must make observations in or through the atmosphere or who are interested in the structure of the atmosphere itself.

As is evident from the title, the author has singled out turbulence and turbulent processes for detailed analysis; and, though there may be some disagreement with the implications of this restriction, a thorough study of turbulent mechanisms and of the relation they bear to wave propagation phenomena is an important and valuable contribution.

Structure function methods and corresponding spectral analysis are employed throughout, as are the concepts of statistical turbulence theory. In fact, on the basis of referenced evidence, almost all of which is of Russian origin and which it is claimed provides convincing substantiation of the Kolmogoroff theory of locally isotropic turbulence, the author proceeds to relate each wave propagation process to the isotropic irregularity structure that he associates with such turbulence.

The book is divided into four parts: I. a review of statistical turbulence theory; II. scattering of electromagnetic and acoustic waves; III. detailed analysis of line-of-sight propagation, both by the method of geometrical optics and by the method of small perturbations; and IV. a comparison of theoretical results and experimental data. In Part II the standard single-scattering (Born) approximation is employed to calculate the scattered fields, which are expressed in terms of the spectral representation of the refractivity irregularities. Similarly in Part III solutions are developed for the usual logarithmic amplitude and phase functions and the intensity and structure of these variables are, in turn, evaluated through the spectral distribution of the refractivity. In both parts the results are generalized to include the important case of locally homogeneous fields.

Although Parts II and III contain much useful information and many important results, in the eyes of this reviewer it is Parts I and IV that give this book its true value. The detailed introduction to the theory of homogeneous and locally homogeneous random fields, the review of the statistical theory of turbulence and the discussion of its role in the study of atmospheric processes, and the development of a model for the turbulent mixing of passive, scalar quantities, such as moisture vapor, all of which are included in Part I, will be especially helpful to the worker with little or no background in turbulence. The empirical results presented and analyzed in Part IV, pertaining both to wind and temperature fluctuations in the atmosphere and to the structure of received acoustic and electromagnetic fields in actual propagation experiments, will provide the investigator with valuable insight that no amount of purely theoretical analysis could supplant.

In spite of its many strong features, this work is open to criticism for the manner in which problems of atmospheric turbulence have been oversimplified. The author appears to consider Kolmogoroff's theory of locally isotropic turbulence the panacea for all problems in which turbulence plays a role. He does not seem squarely to face the fact that, in the analysis of any real situation, one must clearly and carefully relate the small scale structure, which may perhaps be accurately described by that model, to the large scale anisotropic structure, which certainly cannot be properly represented in that way. He criticizes other authors for using phenomenological models of turbulence processes yet falls back on them himself, without comment, when in Chapter 10 he finally attempts to relate the large and small scale structures. One does not find fault with the author's analysis but rather with the manner in which he has approached it and with the impression he has thereby created that all the pertinent turbulence problems are solved by the Kolmogoroff theory, and that all that remains is the application of the results to various wave propagation situations.

This review could not close without brief comment on the translation. Dr. Silverman and the publisher have done another fine job in transcribing this work into English. It reads clearly and smoothly, a number of helpful explanatory comments have been added, and a minimum number of mechanical and editorial errors have been allowed to survive to final production. All in all, this is a valuable contribution, especially since it makes this otherwise inaccessible material, and an extensive bibliography on the subject, readily available to the Western scientific community.

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Semiconductor Devices and Applications, by R. A. Greiner

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. 405 pages+5 index pages+xiv pages+12 references pages+43 appendix pages+5 pages postface. Illus. 6×9¼. $12.50.

This is another in the growing list of books which attempt to cover both semiconductor device principles and circuit applications. The book is almost equally divided between these two areas. The first half is devoted to a review of modern physics, semiconductors, conduction in semiconductors, semiconductor materials preparation, contacts, junctions, diodes, and transistors; the latter half is devoted to small signal and power amplifiers, dc amplifiers, oscillators, and switching and logic circuits. The treatment is roughly at the junior or senior level and is commendable in that the author attempts to interject device understanding into circuit understanding. The author's writing style is facile and his book easy to read.

Relative to the large number of very comprehensive books already available on semiconductors, semiconductor devices, and circuit applications of semiconductor devices, the book under review possesses a number of shortcomings. The first chapter on modern physics is too brief to be worth much to a student unacquainted with modern physics, and is unnecessary for the properly prepared student. In terms of modern device practice, a chapter entirely devoted to metal-semiconductor contacts is wasted and takes space needed for other topics. A number of statements in the first half of the book are either in error or give the wrong emphasis. For example, silicon has greater importance than germanium but not necessarily or completely because of its greater abundance and higher operating temperature; with silicon becoming more widely used than germanium, the most common method to measure lifetime is not the one stated; and a diffused impurity profile is not generally exponential. In places throughout the book strange terminology (strange to a student) is introduced without explanation or any references. Some concepts or formulas are presented without explanation or references and cannot but confuse the student or make him wonder why he doesn't grasp "the obvious." Unfortunately, many of the more powerful concepts introduced by Shockley in analyzing junctions are not included and apparently must be supplied by the classroom instructor. As mentioned, considerable space is given to metal-semiconductor contacts, and important matters such as minority carrier lifetime are not fully developed. In spite of the generality of the title, hardly any devices besides junction diodes and transistors are considered in the text proper.

Despite the limitations of this book, it should be found useful as an aid in instructing students unfamiliar with semiconductor devices and who quite likely intend to specialize in circuit applications. The experienced instructor will know what to add, what to delete, and what to correct, including misspellings of the well-known names Stirling and Hankel.

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Philosophical Impact of Contemporary Physics, by Milič Čapek

Published (1961) by D. Van Nostrand Co., Inc., 120 Alexander St., Princeton, N. J. 399 pages+15 index pages+vxii pages. 6×9¼. $7.50.

The primary purpose of this book is to bring into full focus the contrast between the classical and modern conceptual frameworks of physics. The book begins with a systematic examination of the fundamental classical concepts of space, time, matter, motion, and causality. These concepts and their mutual relations are discussed, and their connection with the philosophical thought of the period are expounded. The radical differences between the new concepts and their classical counterparts are explained, and misinterpretations due to the survival of the classical habits of thought are criticized.

The author of this book is well versed in both philosophy and physics, and is eminently qualified for the rather difficult interdisciplinary task which he has set himself. Currently, Professor of Philosophy at Carleton College in Northfield, Minn., the author formerly taught physics at Doane College, the University of Nebraska, and the University of Olmutz (Olomouc) in Czechoslovakia.

In principle, this book is intended both for students of philosophy interested in learning more about physics, and students of physics interested in gaining a deeper appreciation of the philosophical implications of their subject. In practice, the book will probably appeal more to the former group than to the latter, since it is written with a strong philosophical orientation. While the present reviewer found this book most worthwhile, he did not find it as incisive or illuminating as other books on the same general subject, such as Margenau's "The Nature of Physical Reality." Nevertheless, he feels the present author has made a significant contribution to the borderline area of physics and philosophy.

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Microwave Ferrites, by P. J. B. Clarricoats

Published (1961) by John Wiley and Sons, Inc., 440 Park Avenue S., New York 16, N. Y. 253 pages+4 index pages+xi pages+references by chapter. Illus. 6¼×10. $8.00.

This book contains an introductory chapter on general properties of ferrites, chapters on microwave behavior of ferrites and loss mechanisms in ferrites, a long chapter on propagation in media containing ferrites and a chapter describing briefly some microwave ferrite devices. The material in the first three chapters mentioned above may well be the most useful, although about forty percent of the book is the propagation chapter which discusses the boundary value problems involved in various configurations. It is good to have this material brought together in one place with a consistent notation, although the components designers for whom the book is intended have generally used empirical methods because of the complexity of the analytical results. In this connection, the use of mathematical appendixes makes the text of the book relatively easy to read.

In a book such as this, one might have hoped to find a discussion of nonlinear effects, but the subject is mentioned only briefly in connection with loss mechanisms.

Workers in the field are forced to adapt themselves to a mixture of cgs units and mks units, and the author has chosen to go along with the usual mixture. Two chapters on ferrite properties use cgs units, and the rest of the book uses mks units except that magnetic field strength is quoted in oersteds, and occasionally such terms as db per inch are used.

This book is definitely not intended as a designer's handbook. The emphasis on the analytical results makes it rather specialized. However, the exposition is clear and workers in the area of microwave ferrites should look at the book and may well want to own it.

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Foundation for Electric Network Theory, by Myril B. Reed

Published (1961) by Prentice-Hall, Inc., Englewood Cliffs, N. J. 323 pages+5 index pages+xii pages+28 appendix pages. Illus. 6¼×9¼. $13.00.

Professor Reed approaches resistive networks by way of linear graph theory. He states in the preface: " . . . this book may be used as a beginning course at the sophomore level or as an elective junior or senior course. It can also be used as supplementary or reference material for senior courses, graduate courses, and self or formal training of engineers in industry."

The first major division of the book, Chapters 2 through 8, deals with linear graph theory as the foundation for formulating a set of linear simultaneous equations relating the voltage variables and current variables in a graph by Kirchhoff's laws. The second major part, Chapters 10 through 14, uses the results of the first part to derive properties of resistive networks. Chapter 1 gives some introductory material; Chapter 9 discusses such topics as energy, power, rms, and average; and the appendix presents an introduction to matrix algebra.

Within the domain of material selected, the arrangement is logical and, with the help of many illustrative examples, the explanations are very clear. There are many problems at the end of every chapter, including the appendix, to enable the student to digest the material.

Although Professor Reed claims to operate from a "near-rigorous basis," so as to give the students the solidest possible foundation, many theorems are listed as properties and presented without proof, on the ground that a "working knowledge" is being presented. In the reviewer's opinion, this leaves the students with many unexplained mysteries. Therefore, it is doubtful whether it really is worthwhile to go into the detailed properties of linear graphs at length, as is done in this book, instead of introducing just a few of the necessary concepts to enable the student to apply Kirchhoff's laws to networks as is done in Professor Guillemin's book "Introductory Circuit Theory." It is true that a student using Professor Guillemin's book has to accept, without a "rigorous" proof, that the set of equations obtained is independent. It is equally true, however, that a student using Professor Reed's book would have to take for granted that the coefficient matrices in equations (6.3.1) and (6.3.2) are non-singular.

There are some minor criticisms:

1) It is rather disappointing that Professor Reed does not include a discussion of node-pair analysis based on cut-set matrices.
2) Because of the number of symbols used in the book, a list of symbols at the end would be very helpful.
3) In the appendix, a matrix is defined as an array of entries of quite general nature, *viz.*, symbols, operators, etc. But later, in the proofs of the various properties of associativity, commutativity, and distributivity of matrix operations, nothing is mentioned about the necessary assumptions on the operations of the individual entries.
4) According to definition 1.11.2, a negative resistor would be classified as a passive element.
5) Although the statements about reciprocity (properties 10.4.1 and 10.4.2) implicitly exclude so-called controlled sources from the network, the nonreciprocal nature of vacuum tube and transistor circuits are not specifically mentioned.
6) It is somewhat confusing that, in Chapter 1, lower case letters a and b are used both to describe meter readings made on an element and to designate the terminals of an element.

In conclusion, it is the opinion of this reviewer that if this book is to be used fruitfully as a sophomore text, much guidance from the instructor is needed.

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Plasma Acceleration—The Fourth Lockheed Symposium on Magnetohydrodynamics, Sidney W Kash, Ed.

Published (1960) by Stanford University Press, Stanford, Calif. 117 pages+vii pages+bibliography by chapter. Illus. 6⅛×9¼. $4.25.

The book is a compilation of nine papers presented at the Fourth Symposium on Magnetohydrodynamics at the Lockheed Missile and Space Division on December 2, 1959. The first two papers discuss steady-state acceleration of plasma by crossed electric and magnetic fields. In one system, the magnetic field is derived by currents in the plasma; and, in the other, it is externally applied. In the third paper are derived scaling relations for devices employing fully ionized gases. In the fourth paper, plasma acceleration by a continuous train of transverse magnetic waves in an electrodeless rectangular duct is investigated. The next four papers deal with the pulsed plasma accelerators which utilize the energy obtained in the discharge of high energy capacitors. The last paper discusses the acceleration of

ionized material by purely electrostatic means; a novel electrode arrangement is presented and problems in the design of the electrostatic accelerator are analyzed.

The book will be of interest to physicists and electronic engineers who are concerned with space vehicle propulsion or laboratory test "tunnels."

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Transmission of Information, by Robert M. Fano

Published (1961) by The Technology Press, Mass. Inst. Tech., Cambridge, and John Wiley and Sons, Inc., 440 Park Ave. South, New York 16, N. Y. 340 pages +3 index pages +x pages +43 appendix pages. Illus. 6×9¼. $7.50.

The author has been active in the field of information theory since its very beginnings and has taught a graduate course on the subject over the past ten years at the Massachusetts Institute of Technology. This book is an outgrowth of his lecture notes.

The scope of the book is indicated by the following list of chapter headings and a brief description of their content. The first chapter, "The Transmission of Information," is in fact an introduction. Communication systems are defined; and related problems, which form the subject of information theory, are outlined. The major result, existence and meaning of the channel capacity, is stated at this point in advance of a proof. Chapter 2, "A Measure of Information," introduces the concepts of mutual information, self-information and, by averaging over ensembles of events, the various entropies. The nomenclature used is in conformance with the IRE Standard on Information Theory.[1] In these Standards, however, "self-information" is also called "information content," and "mutual information" is called "transinformation." In Chapter 3, "Simple Message Ensembles," the encoding of an ensemble of messages is discussed. Optimum codes, which minimize the average length of the coded message, are constructed. In Chapter 4, "Discrete Stochastic Sources," *i.e.*, sequences of messages specified by some statistical properties, are considered. Encoding is used to define their rates. The important case of Markov sources is discussed in some detail. The next chapter, "Transmission Channels," describes and classifies simple models of channels and defines their capacity as the maximum mutual information per event between the input and the output, when the probability of the input events is varied. It is computed for discrete channels and time-discrete continuous channels with additive noise. Gaussian noise and band-limited channels are considered. In Chapter 6, "Channel Encoding and Decoding," it is shown that, for some channels, the capacity defined by means of the mutual information is also the upper limit to the rates of transmission that can be approached with an arbitrary small probability of error. Conversely, a rate of transmission larger than the capacity implies a finite rate of error. The three remaining chapters treat of encoding and decoding for transmission through discrete constant channels. Chapter 7 is "Encoding for Binary Symmetric Channels," and Chapter 9, "Encoding for Discrete Constant Channels." In Chapter 8, "Multinomial Distributions," special mathematical techniques (tilted distributions) are introduced to solve the problems of upper and lower bounds for the probabilities of error in the last chapter.

The text is reproduced by photo-offset. Errors, beside those mentioned in the errata sheet, are very minor ones: in formula 7.63, p. 233, the exponent of $(1-p)$ should be $n-k$, and on page 355, problem 6.8 should be numbered 4.8. The index also is not quite complete since ergodic and tilting, for example, are not listed.

The last chapter contains original material recently developed by the author and some unpublished results of C. E. Shannon.

On the whole, the subject is presented clearly and simply. The book should be an excellent text for a first course on the subject. The mathematical background of the reader does not have to include more than the fundamentals of probability theory and of Fourier analysis. A set of 80 problems should help the student assimilate the material and appreciate some of its practical applications.

The question of the usefulness of information theory is sometimes raised in engineering circles. The author, certainly aware of this, gives an answer in the summary of the first chapter and in the conclusion of the last one. It is, in essence, that "communication engineering is at the threshold of a major technical revolution." In this revolution, data processing will play an important role, and a deep understanding of the foundations of the theory will be necessary. "Transmission of Information" will certainly contribute to this understanding.

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[1] "IRE Standard on Information Theory: Definition of Terms, 1958," PROC. IRE, pp. 1646–1648, September, 1958.

Digital Applications of Magnetic Devices, Albert J. Meyerhoff, Ed.

Published (1960) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 569 pages +14 index pages +xix pages +bibliography by chapter +11 appendix pages +7 glossary pages. Illus. 6×9¼. $14.00.

This book, which could serve either as a text or reference handbook, covers a large number of digital logic and memory circuits whose operation is based on the use of magnetic components or magnetic interactions. Many of the devices presented appear in a book for the first time. The subject matter is, for the most part, covered in a very comprehensive manner.

The first seven chapters of this 604 page book serve as an introduction to the circuits described in subsequent chapters. An elementary introduction to magnetism is given. The properties of magnetic cores, diodes, and transistors are discussed. The parameters and units to be used are defined.

The next fourteen chapters, or approximately one-half of the book, are concerned with core-diode circuits. These circuits have been classified into three basic types. They are: 1) parallel magnetic-pulse amplifiers, 2) delay parallel magnetic-pulse amplifiers, and 3) series magnetic-pulse amplifiers. The treatment is rather extensive and complete.

Memories suitable for high-speed digital systems are described in the following five chapters. Coincident-current and linear-selection memories are described. Following a discussion of the general principles, specific designs are presented. Thin magnetic film, twistor, and superconductive memory elements are treated briefly. Three chapters are then devoted to the design and application of regenerative transistor-core circuits. The final chapters concern carrier magnetic amplifiers and multiaperture core circuits.

With the exception of the introductory material on ferromagnetism and core properties, most of the material presented is covered in sufficient detail and is handled well. The book would have been given better balance if the sections on core-diode and core-transistor circuits were substantially shortened and the introductory material expanded. The use of mirror symbols to simplify core circuit presentations would also have been helpful.

It is understandable that the twenty-five authors would concentrate on material with which they had firsthand experience. The final result, though, has both a lack of continuity and an omission of many useful devices. An example of this would be the parametron which logically belongs in the section on carrier magnetic amplifiers. It is hoped that this will be remedied in future editions.

In spite of the lack of balance, this book will definitely serve a useful purpose in a rapidly growing field.

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Scanning the Transactions

Majority rule, which in various forms governs much of the affairs of man, is also governing the operation of more and more of his machines and equipment. The theory is that several heads, or parts, are often more reliable than one. The use of parallel redundant circuits to improve the reliability of electronic equipment is already a familiar technique. The majority rule idea carries this technique a step further by adding a vote-taking element to the system which polls the outputs of the redundant parts of the system to determine, on the basis of a simple majority vote, which output is correct. Thus, each redundant part assumes more than the role of a stand-by and becomes a full-time performer that has equal billing with its counterpart. It has recently been proposed that the reliability of a system could be further improved if decisions were based on a weighted vote which took into account the past voting record of each redundant part, rather than on a simple majority vote. To accomplish this, an adaptive vote-taker has been proposed which is able to reduce the significance it gives to the outputs from those parts which past experience shows are apt to make mistakes. Thus the vote-taker acts as an automatic repairman by gradually eliminating defective parts from the operations of the system. The heart of the proposed adaptive vote-taker is an element that provides variable gain with memory. A variable resistor with memory, called a memistor, has been successfully applied to this function. It is believed that, with the application of adaptive techniques to redundant systems, the dependability and life expectancy of a system can actually be made to exceed that of its component parts. (B. Widrow, *et al.*, "Birth, life, and death in microelectronic systems," IRE TRANS. ON MILITARY ELECTRONICS, July, 1961.)

The alphabet has played a greater role in the progress of mankind than many of us realize. The ancients who devised the letters of the alphabet deserve more credit for their contributions to many modern-day developments. In the electronics field, for example, it is difficult to envision what life would be like if there were no A-scopes, F-layers, G-string transmission lines, J-scans, X-rays, or z-transforms. Even filters have their Q's. More remarkable, however, is how those living centuries ago anticipated the physical shapes of things to come. It seems clear that in designing the characters of the alphabet they foresaw that today we would need such things as I-beams and S-curves, and would frown on such things as U-turns. Their foresight even carried into electronics where we find such examples as H-guides and magic T's. Whether they foresaw the new surface-wave structure that was recently developed is perhaps open to question. In any event, another worthy member has been added to the alphabetic list which is called a V-line. It is a wedge-shaped structure consisting of two conducting plates which form the sides and a dielectric medium contained within the V which supports the propagation of surface waves. A noteworthy feature is that the application of biasing potentials between the conducting plates may provide a convenient means of electronically controlling the propagation characteristics. The V-line thus makes an interesting addition to the transmission line family. (P. Diament, *et al.*, "A dielectric surface-wave structure: the V-line," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, July, 1961.)

How to supply the L in RLC is posing a problem today. The general trend toward microminiaturization has placed increasing emphasis on the use of semiconductor circuits. Unfortunately, it is difficult to produce useful inductance in semiconductor materials. The alternative of using wire-wound inductors is unattractive because coils are bulky and hard to microminiaturize. How, then, do we expect to obtain microminiature versions of RLC filter circuits? The answer lies in the development of inductive semiconductor elements. Although the going is not easy, some progress has been made in this direction. Forward-biased junction diodes, for example, can be made to behave like inductive elements. Unfortunately, they exhibit a low Q. Coupling them with negative resistance devices has been tried in order to increase the Q, but serious temperature and stability problems have arisen. Better success has been achieved by combining a transistor with a 90° phase-shift network to make the circuit appear inductive to an external source, but again only a moderate Q is obtainable because of the low input impedance of the transistor. A recent evaluation of these and other possibilities indicates that at the present state of the art the most promising approach lies in operating a transistor in the alpha cutoff region to produce an "inductive transistor." The problem is an important one and further progress will be watched with interest. (H. G. Dill, "Inductive semiconductor elements and their application in bandpass amplifiers," IRE TRANS. ON MILITARY ELECTRONICS, July, 1961.)

Abstracts of IRE Transactions

The following issues of TRANSACTIONS have recently been published, and are now available from The Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y., at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	IRE Members	Libraries and Colleges	Non-Members
Aerospace and Navigational Electronics	ANE-8, No. 2	$2.25	$3.25	$4.50
Electron Devices	ED-8, No. 4	2.25	3.25	4.50
Microwave Theory and Techniques	MTT-9. No. 4	2.25	3.25	4.50
Military Electronics	MIL-5, No. 3	2.25	3.25	4.50
Nuclear Science	NS-8, No. 3	2.25	3.25	4.50

Aerospace and Navigational Electronics

VOL. ANE-8, NO. 2, JUNE, 1961

The Editor Reports –(p. 46)
1961 Pioneer Award—(p. 47)

Control Concepts of the AN/GSN-11 Terminal Air Traffic Control System—S. D. Moxley, Jr. (p. 51)

Capabilities of a newly completed terminal-area air traffic control system are discussed, in terms of recent theories of flow control. Adaptability of the system to gain the benefits of early arrival scheduling, time separation, sequencing, schedule resolution, shortest common path, dual and crossing runways is described. The author defends close-control vectoring as the only means to achieving both improved safety and increased traffic rate and discusses the system's philosophy of increasing precision with decreasing range. Getting back to the pilot, means by which the system simplifies approach procedures, reduces congestion aloft, and provides continuous radar tracking are also discussed.

Logarithmic Navigation for Precise Guidance of Space Vehicles—W. G. Green (p. 59)

The principles of logarithmic guidance are derived and their application to various space-flight guidance problems is discussed. Logarithmic guidance is shown to be ideal during the terminus of control where considerations of minimum fuel, minimum heating etc., can be subordinated to precise matching of vehicle kinematics to the desired trajectory. This precise trajectory control is achieved utilizing velocity and position measurements to govern the vehicle deceleration. Multidimensional effects are considered and it is shown that various "degrees of control" of velocity vector magnitude and angle, time of arrival, accelerations, geographical or inertial directions of approach, etc., can be achieved.

The tolerance of logarithmic guidance to instrumentation errors and parameter variations are confirmed by error analyses of these guidance principles applied to the control of the velocity vector during the deceleration process.

Optimization of a Generalized Velocity-Inertial System—W. A. Porter and L. F. Kazda (p. 72)

In this paper the variational calculus techniques of the Wiener-Kolmogoroff optimum filter theory are employed to develop the statistically optimum form of a generalized hybrid velocity-inertial system. A velocity-inertial system is considered which is general in both form and application. The form of the system encompasses pure inertial, pure Doppler, and a large family of Doppler-inertial hybrid systems including the present-day second- and third-order Doppler inertial navigation and stabilization systems. The system may be used for a wide range of applications including any linear combination of acceleration, velocity, or vertical-reference sensing.

The general system form is developed by employing an unspecified filter to mix the inertially derived signal with the signal from the auxiliary velocity sensor. As a result of utilizing this general system form, a single general error equation is found which represents the system error for each of the above system forms and applications. Using a linear analysis and the minimum mean-square error criteria, an optimum system form is found for the complete range of possible system applications.

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Electron Devices

VOL. ED-8, NO. 4, JULY, 1961

On the Heating of Output Windows of Microwave Tubes by Electron Bombardment—D. H. Preist and R. C. Talcott (p. 243)

An account is given of an experimental investigation of the behavior of output windows, the particular type chosen for study being the cylindrical type in a cavity resonator. By feeding microwave power into the cavity and measuring electric field and the heat dissipated in the window and in the metal walls, a nonlinear window-heating effect was discovered above a certain critical field strength. In this condition, watts dissipated at the window vary as the fifth or higher power of the electric field strength and the resulting thermal stresses may easily destroy it. Further experiments were performed on various modifications of the original structure in order to isolate and study the phenomena responsible for the heating. Some of these tests were made using an axial magnetic field.

It was found that there are two distinct mechanisms of heating, one requiring the presence of the magnetic field and the other one not requiring it. Both involve energy exchange between the electric field and the window by free electrons which dance on the window surface synchronously with the alternating field and multiply by a secondary emission or multipactor process above a certain field strength. The critical field strength may be greatly reduced by the presence of contaminating films on the dielectric surface, and may be greatly increased by suitable surface treatment.

It is shown that these effects are not restricted to cylindrical windows but will tend to occur at any dielectric surface in vacuum if the tangential RF field is strong enough.

The Spacistor—J. M. Lavine, W. Rindner, B. Nost, and R. F. Nelson (p. 252)

A brief description is given of several Spacistor structures and the techniques of fabrication. Detailed measurements of the four-pole parameters are reported and described with the aid of physical models. The mutual conductance is shown to depend upon injector-modulator spacing, injector current and modulator bias, with little variation to 25 Mc. Improvement in both g_m and shielding is shown to occur when the modulator is injecting charge. Values of g_m of the order of several hundred micromhos have been observed. The input resistance depends upon the characteristics of the modulator contact in the space-charge region and upon the dc bias. Under forward-biased modulator conditions (optimum g_m) values of input resistance of the order of several hundred thousand ohms are obtained. Under zero-bias or reverse-bias conditions (somewhat reduced g_m), values of several megohms are obtained. Values of output resistance ranging to several hundred thousand ohms have been observed with possibilities of larger values being obtained with improved design. Low frequency voltage gain of the order of 10 and power gain of the order of several hundred have been exhibited. The reduction of power gain at 25 Mc is attributed mostly to the array of contact and device resistances, and device and header capacities, some of which may be reduced by appropriate design. These prevented the elimination of shunting capacities and precluded higher frequency measurements.

Preparation and Properties of Electroluminescent Phosphors for Display Devices—H. F. Ivey and W. A. Thornton (p. 265)

Electroluminescent phosphors for use in some electronic display devices should have brightness-voltage characteristics differing from those used in light sources, for example. The various factors of phosphor preparation and of cell construction which affect these characteristics have been studied. Special phosphor powders with unusually high or unusually low values of "discrimination ratio" (a quantity related to the change in output as the applied voltage is varied) have been developed and their performance described. It has also been shown that thin phosphor films afford extremely high values of discrimination ratio,

particularly if operated on dc. The use of nonlinear materials (especially zinc oxide) as auxiliary layers to increase the discrimination ratio obtained with a given phosphor has been studied.

N-Beam Nonlinear Traveling-Wave Amplifier Analysis—J. E. Rowe (p. 279)

In large-diameter electron stream traveling-wave amplifiers there is a significant variation of the circuit electric field across the stream. Most analyses neglect the effects of these field variations. An approximate method has been used to study the effect of transverse field variations on the gain and efficiency of large-signal traveling-wave amplifiers. The annular electron stream is subdivided into a number of annular rings each containing an equal fraction of the total current. It is assumed that the circuit field varies over the stream cross section and that each segment has a different coupling to the circuit. The space-charge field is assumed independent of radius. The effect on gain and efficiency is calculated for a stream diameter of $B = 1$ and it is found that a subdivision of the stream into two segments with different couplings to the circuit gives a lower saturation efficiency but that further subdivision does not appreciably change the results.

Traveling-Wave Tubes with Ferrite Attenuators—F. K. Mullen (p. 284)

In order to determine the efficacy of ferrite isolators used as internal attenuators in traveling-wave tubes, two TWT's incorporating such isolators were developed, one using orthodox magnetic focusing, the other using electrostatic focusing. The results of the experimentation indicate that the ferrite can, in some cases, offer improvement in gain and efficiency over reciprocal attenuators. The limitations of this device are discussed together with some particular advantages it may offer for special applications.

Design criteria for such isolators and the effects of the ferrite as a circuit element in TWT's, are also investigated.

The Reflex Klystron as a Regenerative Mixer—B. G. Whitford (p. 289)

A simple receiver using a single reflex klystron as a mixer and local oscillator is described. Experimental data show typical performance, and a theory based on reflex-klystron and transmission-line theory is proposed which explains the observed behavior. The S- and X-band tangential sensitivities are about 15 db better than can be realized with a low-level crystal video detector and it is thought that this figure may be exceeded at the higher microwave frequencies where crystal sensitivity is relatively poor.

A Ladder Structure for Millimeter Waves—E. A. Ash and A. C. Studd (p. 294)

A planar structure, closely related to the stub-supported ring and bar structure, is analyzed using Fletcher's method. The computed dispersion characteristics and interaction impedances show that it should prove useful in millimeter generators. Experiments at X band have shown that oscillation bands of over 1.7:1 are possible working on a backward mode. Forward-mode operation with higher output powers, but with a more restricted bandwidth can also be realized.

Some High-Power Window Failures—J. R. M. Vaughan (p. 302)

Two types of RF window failure in a high-power magnetron are investigated. Cracking is shown to depend on electrostatic charging of an evaporated metallic deposit on the window. Punctures were investigated by observations of X-ray stereo-autograms, and are shown to result from an internal multi-pactor discharge, which in turn is enabled to function at high voltages by the stray magnetic field. Methods for curing both types of defect are given, with evidence that they are successful. The methods are applicable to other high-power microwave tubes, both magnetrons and klystrons.

Crossed-Field Electron Interaction of the Distributed-Emission Space-Charge-Limited Type—J. F. Hull (p. 309)

A semitheoretical, semiheuristic analysis is given of crossed-field electron interaction of the distributed-emission type. The emphasis of this paper is on the over-all picture of this complex type of interaction, rather than on mathematical exactness. The results are given in terms of five equations from which can be calculated the amount of dc applied power which is converted to real and reactive RF power under given conditions of frequency, RF potential, and magnetic-field strength. These equations are applied to the magnetron oscillator to predict power delivered to the anode by the electron bunches, frequency-pushing characteristics, V-I characteristics, and Rieke Diagrams. The agreement of the experimental data with the calculations over this wide range of characteristics confirms the validity of the analysis.

Some Experiments with a New Type of Crossed-Field Gun—T. A. Midford and G. S. Kino (p. 344)

The construction and testing of a crossed-field gun based on an exact space-charge flow solution will be described. All the characteristic parameters of a gun of this type may be obtained analytically. Experimentally the gun was found to perform essentially as predicted. This gun has a moderately high convergence and produces a beam whose gross features give a very good approximation to planar Brillouin flow. Work by Anderson has indicated that crossed-field beams produced by a gun of this type may be somewhat less noisy than similar beams produced by other types of guns.

High-Frequency Behavior of the Crossed-Field Potential Minimum—T. Van Duzer and J. R. Whinnery (p. 331)

A time-domain analysis of the high-frequency currents in a crossed-field potential minimum is presented. The distinctions between ordinary and crossed-field potential minimums are discussed and a one-dimensional model for the crossed-field case is described. The general form of the solution is given as an infinite array of coupled integral equations and the required expressions for the coefficients are developed. A reduced array of equations is used for a numerical analysis of a diode with a 100-gauss transverse magnetic field. The numerical analysis revealed only a minor modification of the current fluctuations as a result of the transverse magnetic field, although experiments have shown excess-energy crossed-field currents related to the potential minimum with magnetic fields lower than this.

Contributors—(p. 342)

Microwave Theory and Techniques

Vol. MTT-9, No. 4, July, 1961

Report on Advances in Microwave Theory and Techniques—1960—R. C. Hansen and M. T. Weiss (p. 278)

The Short Pulse Behavior of Lossy Tapered Transmission Lines—R. Stapelfeldt and F. J. Young (p. 290)

An analytic method is given which allows the design engineer to assess rapidly the short pulse characteristics of any given tapered-transmission-line type of pulse transformer. The method allows inclusion of both skin-effect losses and losses which are independent of frequency. The effects of mismatching at either end are shown to be as important as the taper function of the line itself. The results of this approximate method are expressed as simple integrals and matching terms to which it is easy to attach physical significance.

The method is applied to the analysis of two tapered-line pulse transformers which are geometrically uniform coaxial structures with tapered dielectric constants. The line whose nominal characteristic impedance is an exponential function of electrical position is shown to have a good rise time and tilt distortion characteristics.

A Waveguide Quadruplexer System—P. Foldes and T. B. Thomson (p. 297)

Practical design considerations are presented for a relatively simple four-channel waveguide branching system. The most important electrical characteristic of this system is an extremely low pass-band reflection coefficient (better than 1.5 per cent in 30-Mc band) which has only very small variations with the environmental conditions.

Some Recent Findings in Microwave Storage—J. D. Kellett (p. 306)

This paper describes an experimental investigation of frequency memory in a recirculating amplifier storage device. The objective of the investigation was to determine what mechanism caused injected energy to shift to preferred storage frequencies. Using fast-acting crystal switches, the output energy was selectively viewed, and it was found that the energy in any circulation when viewed separately was of the input frequency. The spectrum photographs which are included in the paper show that the recirculating amplifier when operating with an open loop gain greater than unity does not oscillate at preferred frequencies.

Excess Noise in Microwave Detector Diodes—J. J. Faris and J. M. Richardson (p. 312)

The dependence of available excess noise in type 1N26 microwave crystal-diode rectifiers on applied microwave power was measured. This may be approximated by a power law with constants characteristic of the particular crystal. As a consequence of the dependence of both excess noise and dc rectified power on input-power level, there is a level which minimizes the ratio of these quantities. Similarly, in the case of a modulated microwave carrier there is an input level which minimizes the ratio of excess noise to demodulated power, and so provides optimum detection of small modulation.

Quantum Fluctuations in Microwave Radiometry—L. P. Bolgiano, Jr. (p. 315).

This paper assesses the possible significance of the quantum nature of electromagnetic radiation in limiting the measurement accuracy attainable with a microwave radiometer. Analogies are shown to exist between the form of a formula describing fluctuations in the radiometer output, and both a formula describing the radiometer input signal, and also, a formula describing the output of a photocell detector. Detailed quantum mechanical consideration of the processes of amplification and detection are circumvented by considering how the formula for fluctuations in the radiometer output might be modified so as to make it consistent with the measurement precision implied by these other formulas. A modified formula is suggested which includes a quantum fluctuation whose magnitude depends on signal power.

On the Resolution of a Class of Waveguide Discontinuity Problems by the Use of Singular Integral Equations—L. Lewin (p. 321)

It is shown that a considerable number of solutions of rectangular waveguide problems appearing in the literature are all special cases of a general treatment focused around the known solution of a singular integral equation. In terms of this a number of typical results are re-examined. The method is then applied to four new configurations, and the range of application and the limitations are examined.

A Dielectric Surface-Wave Structure: the V-Line—P. Diament, S. P. Schlesinger, and A. Vigants (p. 332)

Properties of the V-line, a wedge-shaped surface-wave structure comprising a cylindrical dielectric binding medium of sectorial cross section supported by two conducting plates, are considered in terms of its higher-order hybrid modes of propagation. Practical modifications of the ideal structure are emphasized.

Design curves and equations are presented to determine various propagation parameters and their significance is discussed. Experimental verification of the theory is described.

Wave Propagation in a Medium with a Progressive Sinusoidal Disturbance—A. Hessel and A. A. Oliner (p. 337)

A recent paper by Simon derives approximate results, employing only three space harmonics, for the propagation characteristics of an electromagnetic wave traveling in a medium possessing a progressive sinusoidal disturbance. A rigorous result is presented here for this same problem, taking into account all of the space harmonics; also, a sufficiency condition for the convergence of this solution is discussed. This sufficiency condition is not satisfied in a particular case treated by Simon. It is shown that his analysis of this case is in error, and that the total field is singular there. The singular nature of the field is associated with "supersonic" effects in the medium containing the progressive disturbance.

A New Broad-Band Absorption Modulator for Rapid Switching of Microwave Power—F. Reggia (p. 343)

This paper describes a new technique for obtaining a broad-band absorption modulator for high-speed switching or amplitude modulation of microwave power. This ferrite modulator, an outgrowth of the longitudinal-field rectangular-waveguide phase shifter, has electrical characteristics particularly desirable in a microwave switch. These include a zero-field insertion loss of approximately 0.5 db in the ON state, an isolation of greater than 60 db in the OFF state which is nearly independent of the magnetic control field in this state, and a nearly matched input impedance for all values of applied field. These electrical characteristics are nearly constant over a 30 per cent bandwidth at X band. Also, it is possible to design the amplitude modulator to have negligible phase shift at the desired operating frequency.

Other characteristics of this ferrite modulator include small physical size, magnetic control fields of less than 50 oersteds, operating temperatures up to 150°C, and a capability of less than one μsec switching time.

Rectangular Waveguide Theoretical CW Average Power Rating—H. E. King (p. 349)

A theoretical CW average power rating, limitation imposed by a temperature rise resulting from power dissipation within the rectangular waveguide walls, can be determined by predicting the rise in temperature. Formulas for the evaluation of the CW average power rating have been developed and are presented here, and the power rating curves are given for the WR-2300 waveguide (320 Mc) through the WR-19 waveguide (60 kMc).

Localized hot spots, associated with a standing wave on a mismatched waveguide, require a derating factor. The axial flow of heat from these high current spots has been considered in calculating and plotting this derating factor.

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PGMTT News—(p. 371)

Military Electronics

VOL. MIL-5, NO. 3, JULY, 1961

Frontispiece—W. L. Doxey (p. 186)

Message from the National Chairman—W. L. Doxey (p. 187)

Frontispiece—J. E. Thomas, Jr. (p. 188)

Editorial—J. E. Thomas, Jr. (p. 189)

Birth, Life and Death in Microelectronic Systems—B. Widrow, W. H. Pierce, and J. B. Angell (p. 191)

In order to exploit the technological promises of microelectronics, electronic system techniques must be developed so that defective portions of a system can be tolerated without system malfunction. Such defects might be introduced during manufacture (at birth), or cause errors during operation (life). The number of permanent failures which could be endured by a system before it fails completely will determine its lifetime (death).

In this paper, an adaptive vote-taker is proposed which compares the outputs of paralleled (redundant) system parts in a binary system and determines the most probable answer based on past performance of the separate parts. Initially, the vote-taker assigns equal significance to each redundant part, and (in a binary system) requires that a simple majority of the parts be correct. With experience, the vote-taker continually reduces the weight (significance) of the outputs from those parts that make mistakes, thereby gradually eliminating the defective parts. Thus the vote-takers (which also may be paralleled if they are unreliable) act as automatic repairmen which delete defective parts of a system. System dependability and life expectancy can be made to exceed the dependability and average life expectancy of the component parts.

The heart of an adaptive vote-taker is an element providing variable gain with memory. A variable resistor with memory (memistor), which uses electrochemical deposition or removal of copper to achieve the variable memory, has been successfully applied to this function.

Use of Passive Redundancy in Electronic Systems—J. J. Suran (p. 202)

Circuit design predicated solely upon the premise of providing circuits with greater and greater immunity against component tolerances at some point results in an increase in catastrophic failure rates. Thus, circuits which are overdesigned to provide maximum protection against drift failures may actually contribute to a lower system reliability than do those circuits which are designed to operate with tighter component tolerances. These considerations lead to the conclusion that as systems grow in complexity or as reliability requirements are significantly increased for current equipment, a point is reached where further improvement in design procedures will not necessarily lead to further increases in system reliability. A way out of this dilemma appears to be the introduction of some form of redundancy on the component or circuit level in order to overcome the effects of the inevitable occurrence of catastrophic component failures. This paper discusses the application of redundancy techniques, of a nonadaptive and passive nature, to electronic circuits and systems.

Power Dissipation in Microelectronic Transmission Circuits—J. D. Meindl (p. 209)

The interrelationship of power dissipation, gain, stability, terminal impedance values, dynamic range and efficiency is investigated for small-signal amplifiers in the middle range of frequencies. Utilizing a novel circuit-design theory which treats a transistor along with its biasing resistors as a single entity, amplifier designs are derived which combine optimum ac performance and minimum dc power dissipation. The product of ac power gain and dc-to-ac efficiency is found to be a useful figure of merit for microelectronic transmission circuits.

A Thermal Design Approach for Solid-State Encapsulated High-Density Computer Circuits—A. E. Rosenberg and T. C. Taylor (p. 216)

This paper considers the thermal problems associated with the design of high component-density encapsulated circuits, constructed with small solid-state components. The thermal resistance to the dissipation of component-generated heat is shown to consist of that of the encapsulating medium, plus that of the external circuit cooling process. Because the external cooling becomes more difficult as the size of an encapsulated circuit is reduced, a method of constructing such circuits is proposed which minimizes the thermal resistance due to the encapsulating medium. This construction makes a large fraction of the allowable component temperature rise available for use in the external heat dissipation process by providing high thermal conductance paths for the transfer of heat from the surfaces of the components to one surface of the circuit structure. Analytical models are developed for the most important heat transfer processes in the proposed circuit structure. The equations based on these models are arranged in a form suitable for design use, and example designs are presented.

Integration of Microcircuitry into Microassemblies—R. A. Gerhold (p. 227)

The necessity to hybridize newer microcircuitry techniques with existing design capabilities to achieve efficient equipment designs is established. An analysis is presented of the net size and weight advantages calculated for several components of a typical weapons system, as first the micromodule and then, progressively, thin-film and solid-circuit techniques are integrated into the system design. The diminishing returns on the introduction of microcircuitry into many areas of the system emphazise the need for high efficiency in the integration of new techniques. The potential capabilities of an advanced type micromodule, which still retains the standardized microelement dimensions and assembly procedures, are explored. A further advanced modular interconnection of microcircuitry wafers into a projected microassembly is then described. 0.002×0.010-inch copper ribbon conductors are welded to the metallized edge terminations of stacked substrate wafers by an electron beam technique. Interconnection requirement is 1600 terminations per square inch. Sample illustrated showed 2000 terminations per square inch, or 80,000 terminations per cubic inch. Termination capability is increased three times, and effective useful module volume is doubled compared to existing micromodule. Details of materials and processes for the micro-connections are described together with a summary of the statistical reliability evaluation. The latter confirms the high reliability objective for this connection.

A Family of Semiconductor Devices for Microelectronic Applications—E. E. Maiden and W. F. Schnepple (p. 233)

A variety of approaches to microminiaturization is now being widely explored by electronic systems designers. Among the approaches taken, continued reliance is being placed upon discrete active components that can be used in module, welded, thin-film, or hybrid types of microcircuits. By the use of discrete elements, maximum circuit flexibility is retained, tight component tolerances are possible, and production shrinkage of complete circuit functions is minimized, as compared with the solid-state circuit approach. Moreover, "throw-away" maintenance costs are low.

A series of microminiature silicon diodes and transistors has been developed and produced for use in applications where stringent size and weight limitations, and high reliability requirements, exist. Through the use of extremely simple mechanical constructions, surface passivation techniques and impervious

glass-like coatings, low cost can be achieved in mass production without sacrificing reliability. By use of appropriate fabrication processes, alloyed or diffused diodes and transistors can be formed in mesa or planar configurations, and an extension to epitaxial structures can be employed. Electrical characteristics are comparable to, or better than those of existing conventional types of diodes and transistors. At 25°C, failure rates below 0.01 per cent per thousand hours have been substantiated.

This paper begins with an outline of the theory underlying surface protection techniques used, describes the construction and characteristics of several devices in the family, and presents test information proving a high degree of reliability.

Inductive Semiconductor Elements and Their Application in Band Pass Amplifiers—H. G. Dill (p. 239)

Filter circuits using wire-wound inductors are hard to microminiaturize because coils are rather bulky. This paper discusses different inductive semiconductor devices which may replace coils where space is a problem.

Forward-biased diodes, properly designed, behave like very lossy inductances. Combining them with negative-resistance devices increases the Q but creates serious temperature and stability problems.

Relatively temperature-stable inductance elements are possible by combination of a phase shift network with a transistor. The principle, well-known in tube circuitry, gives high inductances with only a moderate Q because of the low input impedance of the transistor.

Promising results have been demonstrated with a transistor operating in the α cutoff region. The device is dc stable, and has a moderate temperature sensitivity which might be partly compensated if necessary. Avalanche multiplication is used to reduce the damping resistance of the inductive transistor.

Simple band-pass amplifier circuits are presented in the last section to demonstrate how to use the inductive transistor in practice.

Fabrication of Microminiaturized Core Memories by Plastic Encapsulation Techniques—G. R. Henderson, W. C. Earl and C. G. Kyratzis (p. 250)

Plastic encapsulation of nonlinear ferrites, and the effect of this encapsulation on the magnetic characteristics of the ferrites, is discussed in this paper. This technique provides greater immunity to shock and vibration damage than conventional core frames and provides bit densities in excess of one million per cubic foot.

Chemical deposition and photographic techniques may be used to form a portion of the wiring matrix. Through-hole plating of ferrites having a 15-mil diameter hole is possible.

Small-evaluation memories have been fabricated, using both toroids and transfluxors. Temperature tests on these memories show that encapsulation causes a slight increase in switching speed and a small decrease on output for a given drive current. These changes remain almost constant over the temperature range. No significant changes have been noted in the noise output.

Techniques outlined are presently being used in the fabrication of microminiaturized, nondestructive memories for missile and satellite application.

MIST Module Electronics—I. Maloff and V. Lally (p. 256)

The MIST module is a proposed building block for electronic telemetry systems for use with weather balloons, providing a minimum hazard to fast-flying aircraft. This electronic telemetry system may be spread in two dimensions while having a minimum build-up or structure in the third dimension. MInimum STructure modules (MIST modules) are a proposed answer to such a requirement. MIST modules have been built on an experimental basis and the brief experimental experience with working modules indicates that they will meet the requirements and provide electronic systems which can be safely floated in the air lanes.

Contributors—(p. 260)

Nuclear Science

Vol. NS-8, No. 3, July, 1961

A Digital Start-up Control Unit for Nuclear Reactors—J. D. Schmidt, B. K. Eriksen, and W. Peil (p. 1)

The use of digital circuits for control of a nuclear reactor in the start-up range was proposed in a paper given by S. N. Lehr and V. P. Mathes at the Nuclear Engineering and Science Conference in 1959. A control of this type eliminates the many periodic adjustments that are required with linear circuits and results in a more accurate, reliable, and maintenance free unit. This paper describes the design and development of such a unit which computes and displays digitally the reactor level and period over nine decades of power. The level of the reactor is obtained from a discriminator and binary counter which resolve 2.5 m.v. pulses spaced as little as one microsecond apart. A time modulation technique incorporating binary counters for comparison and read-out is used to compute the period with a short term memory employed to provide statistical smoothing.

Use of Entrance Hodoscope for Particle Identification in Very-High-Energy Bubble Chamber Experiments—W. Selove, H. Brody, E. Leboy and R. Fullwood (p. 13)

At energies of 10 Bev and higher, it is not easy to obtain physically separated beams of π and K mesons. We have constructed a hodoscope system which will register the identity of of each of 20 particles entering a bubble chamber over a 100 microsecond time interval. Particles are localized by a scintillator hodoscope with matrix elements 1 cm square. Particle identity is determined with Cerenkov counters. The combined information is displayed on an oscilloscope and photographed at each bubble chamber expansion.

The Log Count Rate Period Meter Used with Safety Circuits—H. Christensen and R. B. Stanfield (p. 22)

Equations are presented that give the probable rate of spurious reactor scrams due to counter noise in the period circuits commonly used at low reactor power. It is shown how a filter introduced to reduce the spurious scram rate, will delay the circuit response in the case of true emergencies. A numerical example shows that a compromise can be reached that gives a very satisfactory circuit.

Proportional Control and Pressurizer Parameter Study of a Pressure Tube Reactor—D. E. Rathbone (p. 27)

An investigation of the transient behavior of a pressure tube reactor indicated a definite need for external control to handle even normal load perturbations. Three proportional controllers were developed which were effective in containing the transients of the system for changes in load. A comparison of the three controllers indicated little preference between them with regard to variations in reactor power, primary loop temperature or system pressure. The final selection of a controller was thus an economic consideration. The result of the parameter study of the pressurizer indicated that a pressurizer size of 30 cubic feet, $\frac{1}{8}$ of the total volume of the primary coolant, was adequate for the containment of the pressure and volume surge transients.

Electronic Parts in a Hypernuclear Environment—L. B. Gardner and A. B. Kaufman (p. 35)

A Double-Delay-Line Clipped Linear Amplifier—R. L. Chase and V. Svelto (p. 45)

A compact transistorized linear amplifier has been designed that is suitable for many radiation counting applications. The amplifier delivers symmetrical double-delay-line differentiated output pulses, up to ± 10 volts in amplitude with a differential non-linearity of $\pm 1\%$. It tolerates input signals 400 times full scale without producing spurious output pulses. Both a prompt output and one delayed by two microseconds are available. The clipping lines and the signal delay line are all terminated at both ends so that physically small, relatively imperfect delay lines can be employed. Five amplifiers occupy only $12\frac{1}{4}$ inches in a standard 19 inch relay rack.

A Digital Data Handler for Pulse Particle Accelerators—W. A. Higinbotham and D. W. Potter (p. 51)

This paper described a digital data handler for storing information in a magnetic core memory during the pulse of a synchroton and transferring it to a slow memory during the dead time. Digital measurements of trajectory, pulse height, time of flight, run number, etc., are typical data. The magnetic core memory provides capacity for storing 32 words (events) of 96 bits during a burst. The information contained in the core memory is then transferred to one inch magnetic tape during the dead time of the accelerator. Thence the information may be fed to a computer for future study. For economy, the data handler has one buffer which serves as the input, output and shift register.

Contributors—(p. 57)

Eighth Annual PGNS Meeting—(p. 60)

Abstracts and References

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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

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The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

UDC NUMBERS

Certain changes and extensions in UDC numbers, as published in PE Notes up to an including PE 666, will be introduced in this and subsequent issues. The main changes are:

Artificial satellites:	551.507.362.2	(PE 657)
Semiconductor devices:	621.382	(PE 657)
Velocity-control tubes, klystrons, etc.:	621.385.6	(PE 634)
Quality of received signal, propagation conditions, etc.:	621.391.8	(PE 651)
Color television:	621.397.132	(PE 650)

The "Extensions and Corrections to the UDC," Ser. 3, No. 6, August, 1959, contains details of PE Notes 598–658. This and other UDC publications, including individual PE Notes, are obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands or from The British Standards Institution, 2 Park Street, London, W. 1, England.

ACOUSTICS AND AUDIO FREQUENCIES

534.112-8:621.374.5 2827

Mode Coupling Occurring in the Propagation of Elastic Pulses in Wires—A. H. Meitzler. (*J. Acoust. Soc. Am.*, vol. 33, pp. 435–445; April, 1961.) Longitudinal and flexural ultrasonic pulses in a stretched wire show considerable distortion because of mode coupling at certain critical frequencies. Phase velocity calculations based on the real roots of the Pochhammer equations can be used to predict these critical frequencies for any isotropic elastic material.

A list of organizations which have available English translations of Russian journals in the electronics and allied fields appears each June and December at the end of the Abstracts and References section.

The Index to the Abstracts and References published in the PROC. IRE from February, 1960 through January, 1961 is published by the PROC. IRE, May, 1961, Part II. It is also published by *Electronic Technology* and included in the March, 1961 issue of that Journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

534.2(204) 2828

Long-Range Sound Propagation in the Deep Ocean—F. E. Hale. (*J. Acoust. Soc. Am.*, vol. 33, pp. 456–464; April, 1961.) Information available from laboratory reports and unclassified literature on long-range convergence zone-propagation is discussed.

534.231 2829

Acoustic Intensity Anomalies introduced by Constant Velocity Gradients—M. A. Pedersen. (*J. Acoust. Soc. Am.*, vol. 33, pp. 465–474; April, 1961.) Spurious caustics can arise in ray-theory calculations of intensity based on constant-gradient layers. The use of curved-line segments preserving continuity of velocity and slope in the profile approximation is considered.

534.232:537.228.1 2830

Transient Performance of a Piezoelectric Transducer—M. Redwood. (*J. Acoust. Soc. Am.*, vol. 33, pp. 527–536; April, 1961.) Electrical and mechanical outputs due to step input signals are discussed and analyses made with reference to the exact transmission-line equivalent circuit of a transducer.

534.232-8 2831

Hypersonic Resonance of Quartz at 3500 Mc/s—J. L. Stewart and E. S. Stewart. (*J. Acoust. Soc. Am.*, vol. 33, p. 538; April, 1961.) Pure thickness-mode oscillations in the range 2950–3600 Mc were observed in X-cut crystals 0.1 mm thick. The damping effect of a thin layer of water was detected.

534.232.089.6 2832

Cylindrical-Wave Reciprocity Parameter—R. J. Bobber and G. A. Sabin. (*J. Acoust. Soc. Am.*, vol. 33, pp. 446–451; April, 1961.) A parameter $J_c=(2/\rho c)(R\lambda)^{\frac{1}{2}}L$ for use in standard reciprocity calibration is derived from network theory and from wave propagation theory and its validity is confirmed by experiment.

534.522.1 2833

Diffraction of Light by Very-High-Frequency Ultrasonic Waves—B. R. Rao and J. S. Murty. (*Proc. Phys. Soc. (London)*, vol. 77, pp. 958–964; May 1, 1961.) The investigation was carried out using ultrasonic waves in water in the range 50–230 Mc.

534.846 2834

The Problem of Acoustic Field Conditions in Recording Studios and their Representation in Rooms used for Reproduction—L. Keibs. (*Tech. Mitt. BRF, Berlin*, vol. 4, pp. 125–129; December, 1960.) Theoretical investigations of the relation between direct and indirect sound at different time intervals in rectangular rooms of volume 10,000 m^3 and 100 m^3 where sound is recorded are the basis of considerations of how these conditions can best be reproduced.

534.88:534.417 2835

Comparison Analysis on Four Directional-Receiver Correlators—M. J. Jacobson and R. J. Talham. (*J. Acoust. Soc. Am* , vol. 33, pp. 518–526· April, 1961.) Four systems are analyzed and their SNR's are computed and compared. An examination is also made of their main lobes for various steering directions.

621.395.61.089.6 2836

Reciprocity Calibration of Microphones in a Diffuse Sound Field—H. G. Diestel. (*J. Acoust. Soc. Am.*, vol. 33, pp. 514–518; April, 1961.) The reciprocity parameter for a diffuse sound field follows from that for free space by replacing the distance by the "diffuse-field distance" of a point source. Hence the diffuse-field voltage response can be derived. See 1397 of May.

ANTENNAS AND TRANSMISSION LINES

621.315.212:621.372.2 2837

On the Theory of the Coaxial Cable—P. Poincelot. (*C.R. Acad. Sci. (Paris)*, vol. 251, pp. 1623–1624; October 17, 1960.) Propagation along a coaxial line is studied in terms of Maxwell's equations without introducing transmission-line constants. The method is applicable to all symmetric coaxial structures.

621.315.212:621.372.852.32 2838

A Magnetically Variable Coaxial Attenuator with Low Basic Attenuation—J. Deutsch and M. Offner. (*Frequenz*, vol. 14, pp. 344–347; October, 1960.) The type of attenuator considered uses a gyromagnetic material with an applied dc magnetic field which is variable. In the range 6–8 Gc, such an attenuator can provide maximum attenuation >12 db and a minimum of <0.9 db.

621.315.212.1:621.397.743 2839

Recent Developments in Coaxial Cables for Television Distribution—J. D. S. Hinchliffe. (*J. Brit. IRE*, vol. 21, pp. 457–462; May, 1961.) Constructional improvements incorporated in newly developed cables are discussed. Details are given of methods of testing, and cable characteristics are compared.

621.372.2:621.373.421 2840
Tapered Distributed *RC* Lines or Phase-Shift Oscillators—W. A. Edson. (PROC. IRE, vol. 49, pp. 1021–1024; June, 1961.) The behavior of tapered transmission lines of distributed capacitance and inductance is examined, and summarized in the form of a set of normalized curves.

621.372.8.029.65 2841
Waveguide Equipment for 2-mm Microwaves: Parts 1 and 2—C. W. van Es, M. Gevers, and F. C. de Ronde. (*Philips Tech. Rev.*, vol. 22, pp. 113–125 and 181–189; February 16 and March 14, 1961.)

621.372.821 2842
Design Notes for Strip-Transmission-Line Tuners—G. T. Orefice. (*Electronic Ind.*, vol. 20, pp. 104–105; March 1961.) Equations are given for the design of a shorted-line piston-capacitor tuner. They are applied to the case of a resonator with a frequency range of 500–1000 mc.

621.372.823:621.372.831.1 2843
The Excitation of Higher Modes by the Fundamental at an Off-Set Junction of Two Circular Waveguides—K. Schnetzler. (*Arch. elekt. Übertragung*, vol. 14, pp. 421–424; October, 1960.) The mode conversion occurring at slightly displaced flange couplings is calculated using formulas given by Solymar (1111 of 1960) with some corrections.

621.372.823:621.396.65 2844
Circular Waveguide for Wide-Band Radio-Link Systems—E. Gillitzer and R. Herz. (*Frequenz*, vol. 14, pp. 347–357; October, 1960.) The properties and characteristics of circular waveguide are discussed and compared with those of rectangular waveguide, with particular reference to their use as antenna feeders in 4-Gc radio-link installations.

621.372.826 2845
TwoWire- Millimetre-Wave Surface Transmission—K. Ishii J. B. Y. Tsui, J. D. Horgan. (PROC. IRE vol. 49, p. 1076; June, 1961.) Experimental results are given relating to propagation on a two-wire line at 58 Gc. Attenuation was much reduced compared with a single-wire line; the spread of the fields appeared to be much smaller.

621.372.831 2846
The Excitation of Higher Modes at Waveguide Junctions—K. Schnetzler. (*Arch. elekt. Übertragung*, vol. 14, pp. 425–431; October, 1960.) Formulas are derived for calculating the spurious modes generated in a gradual transition between different waveguide, including the case in which the spurious mode is propagated only in part of the transition. See also 1726 of June and 2843 above.

621.372.852.1 2847
Design Relations for Resonant Post Waveguide Filters—R. D. Wanselow. (*J. Franklin Inst.*, vol. 271, pp. 94–107; February, 1961.) The procedures and graphs given for the design of waveguide filters with *Q*-factors under load ranging from 4 to 200 are shown to be adequate for most engineering applications.

621.372.852.21 2848
A Microwave Phase-Shifter with Dielectric Prisms—R. Tremblay. (*Canad. J. Phys.*, vol. 39, pp. 409–418; March, 1961. In French.) Two identical prisms in opposition form a dielectric plate in a rectangular waveguide. The displacement of one relative to the other along the hypotenuse changes the effective length of the plate.

621.372.852.323 2849
Resonance Isolators for Millimetre Waves—H. G. Beljers. (*Philips Tech. Rev.*, vol. 22, pp. 11–15; November 4, 1960.) Compact isolators for 35 and 70 Gc using crystal-oriented anisotropic ferrites are described.

621.396.67:624.97.042 2850
Load Specification for Aerial Supports—F. Staiger. (*Rundfunktech. Mitt.*, vol. 4, pp. 253–257; December, 1960.) The problem in load calculations of allowing for ice formation on antenna masts is considered in order to achieve greater uniformity in design specifications.

621.396.67.095.1 2851
Generating a Rotating Polarization—S. R. Penstone and A. W. Adey. (PROC. IRE, vol. 49, p. 1089; June, 1961.) A method is described of generating an electromotive wave whose plane of polarization rotates at 80 cps, for use in rocket determinations of the electron density in the ionosphere. See also 2615 of 1960 (Allen).

621.396.677.029.6 2852
Investigations on Endfire Aerials in the Microwave Region—T. Heller. (*Nachrtech. Z.*, vol. 13, pp. 529–533; November, 1960.) The antennas considered are dielectric-rod radiators of helical-line and Yagi structure, and the zig-zag antenna [2848 of 1959 (Sengupta)]. The Hansen-Woodyard conditions for optimum directivity are determined, and practical measures for improving antenna efficiency are mentioned.

621.396.677.3:621.396.43:551.507.362.2 2853
Communications Satellites using Arrays—M. H. Hansen. (See 3189.)

621.396.677.3:621.397.61 2854
Dipole Arrays for Transmitter Antennas for Television Bands IV/V—H. Laub. (*Frequenz*, vol. 14, pp. 327–334; October, 1960.) Problems of electrical and mechanical design of antenna arrays for the frequency range 470–790 Mc are discussed; a number of arrays are illustrated and examples of polar diagrams are given.

621.396.677.71 2855
Hybrid H-Guide Feeds Flush-Mounted Antennas—G. N. Voronoff. (*Electronics*, vol. 34, pp. 54–57; March 31, 1961.) A two-dimensional array of slot radiators can be excited in this way.

621.396.677.73 2856
Horn Radiators as Gain Standards—K. E. Müller and J. Göller. (*Nachrtech.*, vol. 10, pp. 414 422; September, 1960.) Approximation formulas for gain and field strength are given and the design procedure is summarized, with a comparison of calculated and measured characteristics.

621.396.677.73:621.396.663 2857
Direction-Finding Aerials with High Bearing Accuracy and Wide Frequency Ranges—K. P. Lensch. (*Z. angew. Phys.*, vol. 12, pp. 557 -567; December, 1960.) Report of measurements on conical radiators of differing shape and construction, in the frequency range 100–2400 Mc. Measured radiation patterns are compared with those based on calculated values; agreement is only approximate.

621.397.677.85 2858
A Microwave Lens for Rapid Scanning—A. M. Scheggi and G. Toraldo di Francia. (*Alta Frequenza*, vol. 29, pp. 438–449; October, 1960.) Description of the design and construction of a configuration lens for 360° scanning, and also of a stack of such lenses. The radiation patterns and wavefronts obtained conform to theoretical predictions.

AUTOMATIC COMPUTERS

681.142 2859
The Physics of Computer Elements—C. N. W. Litting. (*Brit. J. Appl. Phys.*, vol. 12, pp. 207–213; May. 1961.) The general principles of acoustic, electrostatic, magnetic, optical, and superconducting storage systems are discussed. Magnetic or superconducting devices show the greatest promise for future high-speed applications.

681.142 2860
Control-Pulse Generation for a Digital Differential Analyser—P. L. Owen, M. F. Partridge, and T. R. H. Sizer. (*Electronic Eng.*, vol. 33, pp. 364–371; June, 1961.) Circuit details are given of a transistorized unit which generates control pulses by a pulse-division method involving some unusual processes.

681.142:539.23 2861
Thin Films as Elements for Automatic Computers—W. E. Proebster. (*Elektrotech. Z., Edn. A.*, vol. 81, pp. 913–920; December 5, 1960.) A review is given covering the application of thin-film storage devices, including the cryotron, to data-processing techniques. Thirty-three references.

681.142:621.382.23 2862
High-Speed Logic Circuits using Diode Logical Networks and Current-Switching Amplifiers—R. J. Miles. (*Mullard Tech. Commun.*, vol. 5, pp. 286–294; March, 1961.) The use of diode networks as an alternative to transistors in logical networks is considered. The circuits are versatile and economical, but require associated amplifiers and are likely to have a limiting switching time of about 10 nsec.

681.142:621.382.23 2863
Output-Coupling Networks for Use with Logical Circuits of the Emitter-Current Switching Type—R. J. Miles. (*Mullard Tech. Commun.* vol. 5, pp. 295–300; March, 1961.) Zener-diode coupling networks required to transfer the collector-voltage swing in emitter-current switching circuits to earth level are considered. The disadvantage of low-coupling efficiency of passive output coupling networks is considered in comparison with the versatility which they give to standard logical elements.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.004.6 2864
Reliability of Electronic Components—(*Nachrtech. Z.*, vol. 13, pp. 505–528; November, 1960.) Four papers based on contributions to a discussion meeting on circuit components and materials held by the Nachrichtentechnische Gesellschaft in Stuttgart, May 5–6, 1960.

1) **Technological Measures for Improving the Reliability of Components**—K. H. J. Rottgardt (pp. 505–512).

2) **Lifetime Investigations on Capacitors**—W. Ackmann (pp. 513–518).

3) **The Effect of Humidity on the Electrical Characteristics of Capacitors**—H. Veith (pp. 519–523).

4) **The Unreliability of Electronic Equipment and its Causes**—H. J. Frundt (pp. 524–528).

621.3.049.7 2865
Recent Advances in Microminiaturization, Reliable Components and Cooling Techniques—G. W. A. Dummer. (*Solid-State Electronics*, vol. 2, pp. 18–34; January, 1961.)

621.3.049.75:621.382 2866
Silicon Epitaxial Microcircuit—W. Glendinning, C. Marlett, J. Allegretti, and D. Shombert. (PROC. IRE, vol. 49, pp. 1087–1088; June, 1961.) A Si microcircuit formed by vapor-phase deposition techniques is described. An eight-layered epitaxial structure was formed which

was capable of oscillation in different modes. Shunting capacitive and resistive leakage effects are discussed.

621.314.224.029.4 2867
The Design and Performance of High-Precision Audio-Frequency Current Transformers—J. J. Hill and A. P. Miller. (*Proc. IEE*, pt. B, vol. 108, pp. 327–332; May, 1961. Discussion, pp. 337–338.) Conventional multilayer current transformers are found to have large high-frequency errors. Details are given of precision single-layer transformers which cover the range 50 cps–30 kc.

621.316.5 2868
Application of Boolean Notation to the Maintenance of Switching Circuits—S. Alexander. (*Electronic Eng.*, vol. 33, pp. 372–374; June, 1961.) Concise notation by Boolean algebra of the operations in the maintenance schedules of complex switching circuits is used to simplify testing and fault location.

621.318.132 2869
High-Stability Ferrite Pot Cores—W. A. Everden. (*J. Brit. IRE*, vol. 21, pp. 409–414; May, 1961.) Describes a range of assemblies for laboratory use and for bulk production of stable inductors.

621.318.57:538.221:621.318.134 2870
Analysis of Ferrite-Core Switching for Practical Applications—P. A. Neeteson. (*Electronic Applic.*, vol. 20, pp. 133–152; February, 1961.) An analytical expression describing the switching process is derived which is suitable for practical circuit design. Measurements confirm the results of calculations.

621.318.57:621.375.3 2871
Core Matrix Driver employing Magnetic Amplifiers—J. Yamato and Y. Suzuki. (*Rev. Elect. Commun. Lab., Japan*, vol. 9, pp. 43–49; January/February, 1961.) A high-frequency magnetic amplifier with high efficiency, good SNR and high-output power has been developed.

621.319.4:621.372.54 2872
Low-Impedance Capacitor Design with Special Application to Filters—C. W. McCutchen and I. D. Howard. (*Electronic Eng.*, vol. 33, pp. 349–353; June, 1961.) Methods of reducing series inductance of conventional capacitors are investigated. Measurements of capacitor characteristics are discussed and an improved method of connection for filter capacitors is described.

621.372.44 2873
The Reactive Behaviour of Nonlinear Negative Differential Resistors—V. Cimagalli. (*Alta Frequenza*, vol. 29, pp. 548–554; October, 1960.) Nonlinear negative-resistance two-poles may have a reactive component when forming part of a given relaxation oscillator circuit

621.372.44 2874
A General Steady-State Analysis of Power /Frequency Relations in Time-Varying Reactances—R. G. Smart. (Proc. IRE, vol. 49, pp. 1051–1058; June, 1961.) This analysis produces the Manley-Rowe results as special cases. An "energy coefficient" is derived which specifies the behavior of an individual frequency transformation. The modes in which an individual transformation can operate are described and related to a wide range of varying-reactance devices.

621.372.5:621.382.3:517.945 2875
The Calculus of Deviations applied to Transistor Circuit and Network Analysis—T. R. Nisbet and W. W. Happ. *J. Brit. IRE*, vol. 21, pp. 437–450; May, 1961.) Network analysis is simplified by the method described; rules are given for solving various transistor problems including the distributed-parameter network.

621.372.54 2876
Some Considerations on Optimum Filters—F. Pandarese. (*Note Recensioni Notiz.*, vol. 9, pp. 1043–1051; November/December, 1960.) An explicit formula is derived which gives the pulse response of a filter for signal detection in noise, where the known signal is expressed in the form of a trigonometric polynomial.

621.372.54 2877
The Design of All-Pass, Low-Pass and Band-Pass Filters with a Tchebycheff-Type Approximation of Constant Group Delay—E. Ulbrich and H. Piloty. (*Arch. elekt. Übertragung*, vol. 14, pp. 451–467; October, 1960.)

621.372.54 2878
Replacement of a Cascade of n Unequal Asymmetric Quadripoles of Equal Iterative Impedance by One Quadripole—W. Herzog. (*Nachrtech. Z.*, vol. 13, pp. 475–477; October, 1960.)

621.372.54 2879
The Cascade Connection of Nonreversible Quadripoles—W. Herzog. (*Nachrtech. Z.* vol. 13, pp. 477–478; October, 1960.)

621.372.54:621.372.57 2880
The Synthesis of Active and Passive Filters by the Root Locus Method—A. Susini. (*Alta Frequenza*, vol. 29, pp. 555–573; October, 1960.)

621.372.54:621.375.4 2881
Active Filter Element and its Application to a Fourier Comb—F. T. May and R. A. Dandl. (*Rev. Sci. Instr.*, vol. 32, pp. 387–391; April, 1961.) A theoretical analysis is given and constructional data are presented for a variable-Q variable-frequency band-pass filter using transistor amplifiers. Stable-Q operation is possible over a Q-factor range of 10–250 for frequencies up to 4 kc.

621.372.57:621.317.361.018.756 2882
The Applicability of Active Filters to the Measurements of the Frequency Spectra of Pulses—H. G. Jungmeister. (*Arch. elekt. Übertragung*, vol. 14, pp. 432–434; October, 1960.) The use in frequency analyzers of active filters comprising amplifiers with double-T type RC networks, results in higher Q-values at low frequencies than is possible with stagger-tuned LC circuits. The error assessment given by Seeger and Stäblein (1497 of 1958) is also applicable to this type of circuit.

621.372.6 2883
Synthesis of Transformerless Active n-Port Networks—I. W. Sandberg. (*Bell Sys. Tech. J.*, vol. 40, pp. 761–783; May, 1961.)

621.373.431.1:621.382.23 2884
A Circuit with Several Stable States using Tunnel Diodes—Y. Hazoni. (*Nuclear Instr. Meth.*, vol. 10, pp. 229–230; March, 1961. In French.) The circuit has four stable states, the time required for a change of state being of the order of 10^{-8} sec. It can serve as a fast prestorage device for a multichannel analyzer or as the basic unit of an associated address scaler.

621.374:621.382.23 2885
A Tunnel-Diode Univibrator and Pulse Height Discriminator—Y. Hazoni. (*Nuclear Instr. Meth.*, vol. 10, pp. 231–233; March, 1961.) An adaptation of the tunnel-diode oscillator circuit (2884 above) and its operation in conjunction with a biased backward diode [see 1959 WESCON Convention Record, pt. 3, pp. 9–31 (Lesk, *et al.*)] are described. Applications in pulse shaping and fast coincidence technique are noted.

621.374.3 2886
A Fast Amplitude Discriminator—A. Sarazin, J. Samueli, and G. Bougnot. (*Nuclear Instr. Meth.*, vol. 10 pp. 202–204; March, 1961. In French) The circuit has a discrimination range of 1–30 v, with a threshold stability within ±0.1 v. The output is a positive pulse 1.5 v in amplitude, with rise time <10 nsec and duration of 200 nsec.

621.374.3:621.382.3 2887
New Ways to Trigger Avalanche Pulse Circuits—H. G. Dill. (Proc. IRE, vol. 49, p. 1093; June, 1961.)

621.374.32 2888
Reversible Decimal Counters—J. L. Goldberg. (*Electronic Tech.* vol. 38, pp. 234–245; July, 1961.) Circuit design for both binary and bi-quinary scales of notation is discussed.

621.374.32:621.382.23 2889
Tunnel-Diode Binary Counter Circuit—H. Ur. (Proc. IRE, vol. 49, pp. 1092–1093; June, 1961.)

621.375.127 2890
Investigation of the Single-Ended Push-Pull Amplifier—F. Cervellati and P. Schiaffino. (*Alta Frequenza*, vol. 29, pp. 512–547; October, 1960.) Theoretical and practical design aspects of various versions of this type of AF amplifier are considered.

621.375.4 2891
A Quick Method for Calculating Transistor Amplifier Circuits—R. R. Vierhout and A. J. H. Vendraik. (*Electronic Eng.*, vol. 33, pp. 375–381; June, 1961.) By neglecting the parameter h_{12} calculations on transistor amplifiers can be greatly simplified. This is applied in calculations of feedback, temperature stability and noise.

621.375.4:621.372.632 2892
Calculation of Mixing Processes in Transistor Stages at Low Frequencies on the Basis of the Equivalent Circuit Diagram—J. S. Vogel and M. J. O. Strutt. (*Arch. elekt. Übertragung*, vol. 14, pp. 435–440; October, 1960.) The theory given in 2516 of August is extended to transistor mixer circuits and to distortion effects in them.

621.375.4.018.783 2893
Nonlinear Distortion of Transistorized Amplifiers—H. H. van Abbe and G. C. van Slagmaat. (*Electronic Appl.*, vol. 20, pp. 159–168; February, 1961.) Sources of distortion are examined and methods are suggested for minimizing their effects.

621.375.432 2894
Designing Amplifiers with Nonlinear Feedback—J. C. Looney. (*Electronics*, vol. 34, pp. 46–49; March 31, 1961.) Design criteria for the use of a nonlinear feedback loop with a transistor amplifier are given. Increase of dynamic range and an exponential response can be obtained.

621.375.9 2895
Design Theory of Optimum Negative-Resistance Amplifiers—E. S. Kuh and J. D. Patterson. (Proc. IRE vol. 49 pp. 1043–1050; June 1961.) An examination is made of amplifiers obtained by imbedding a linear active one-port device in arbitrary three-port devices. The optimum gain-bandwidth formula synthesis procedure and design curves are given for this and other amplifier configurations.

621.375.9:538.569.4 2896
Amplification through Stimulated Emission—the Maser—R. A. Smith. (*Brit. J. Appl. Phys.* vol. 12, pp. 197–206; May, 1961.) An account of the physical principles underlying two- and three-level maser action is given, and expressions for noise factor are derived. Recent maser applications in radio astronomy systems are noted.

621.375.9:538.569.4 2897
General Analysis of Optical, Infrared, and Microwave Maser Oscillator Emission—J. R. Singer and S. Wang. (*Phys. Rev. Lett.*, vol. 6, pp. 351–354; April, 1961.) The equations governing coherent emission are generalized. Amplitude modulation of the signal is to be expected and this can be used for communication purposes.

621.375.9:538.569.4:538.222 2898
Cross-Doping Agents for Rutile Masers—P. F. Chester. (*J. Appl. Phys.*, vol. 32, pp. 866–868; May, 1961.) The electron paramagnetic resonance spectra of rutile doped with Ta, Nb and Ce have been examined at liquid-helium temperatures. Ta and Nb appear to be suitable for securing a short spin-lattice relaxation time.

621.375.9:621.372.2 2899
A New Microwave Power Amplifier—P. A. Clavier. (Proc. IRE, vol. 49, pp. 1083–1084; June, 1961.) A method is suggested in which one electrode of a Lecher or coaxial line is made electron-emissive. A signal propagated down the line produces an amplified current in the anode.

621.375.9:621.372.44 2900
Nonreciprocal Parametric Amplifier Circuits—L. D. Baldwin. (Proc. IRE, vol. 49, p. 1075; June, 1961.) Two circuit configurations are given which do not use a circulator but which have similar properties to the single-port reflection amplifier with a three-port circulator.

621.375.9:621.372.44 2901
Staggered Operation of Doubly Resonant Parametric Amplifiers—S. Hamada and H. Mukai. (*Rev. Elect. Commun. Lab., Japan*, vol. 9, pp. 20–25; January/February 1961.) A broad bandwidth is achieved by staggered operation of doubly resonant variable-capacitance parametric amplifiers in the 6-Gc band.

621.375.9:621.372.44:621.318.435 2902
On Parametric Excitation with Current-Saturable Inductances—N. S. Prywes. (*J. Franklin Inst.*, vol. 270, pp. 468 491; December, 1960.) A parametric-amplifier circuit with an abrupt nonlinearity is described. A method of analysis of "relaxation oscillations" leads to a physical understanding of the cycle of oscillations and quantitative relations. The stability problem is also solved.

621.375.9:621.372.44:621.382.23 2903
Minimum Noise Figure of the Variable-Capacitance Amplifier—K. Kurokawa and M. Uenohara. (*Bell Sys. Tech. J.*, vol. 40, pp. 695–722; May, 1961.) A discussion of the minimum noise figure of variable-capacitance diode amplifiers, on the assumption that the series resistance is the only parasitic element. Some experimental figures are given for DSB operation at 6 Gc and universal curves are included which illustrate the effect of changing network parameters and component temperatures.

621.375.9:621.372.44:621.382.23 2904
A Solid-State Analogue to a Travelling-Wave Amplifier—G. H. Heilmeier. (Proc. IRE, vol. 49, pp. 1079–1080; June, 1961.) An attempt to improve the dispersive properties of multidiode parametric amplifiers was made by placing stacks of varactor diodes inside an L-band helix. Bandwidths of 15–25 per cent and and a noise figure of 9.8 db were obtained.

621.375.9:621.372.44:621.385.6 2905
A Low-Noise Microwave Quadripole Amplifier—M. Ashkin. (See 3199.)

621.375.9:621.372.44:621.385.63 2906
Travelling-Wave Analysis of a Class of Parametric Amplifiers Based upon the Hill Equation—Fredricks. (See 3205.)

GENERAL PHYSICS

537.226 2907
Electric Fields and Currents in Dielectrics—É. I. Adirovich. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1410–1422; July, 1960.) A general solution is given for the distribution of electric field and charge-carrier concentration in a dielectric between two metallic contacts when field-emission and thermionic currents flow from the cathode.

537.311.5 2908
Potentials in a Conductor of Varying Cross-Section—R. Jaggi. (*Phys. Rev.*, vol. 122, pp. 448–449; April 15, 1961.) The importance of the eigen-Hall effect [see 1497 of 1954 (Busch and Jaggi)] in masking the Bernoulli voltage in experimental measurements is pointed out.

537.525:538.69:621.314.6 2909
Rectification Effect in Gaseous Discharges in Crossed Magnetic and Electric Fields—G. Boucher and O. Doehler. (*C.R. Acad. Sci., (Paris)*, vol. 251, pp. 59 61; July 4, 1960.) Description of an experimental crossed-field diode consisting of two coaxial electrodes of stainless-steel tube, the inner electrode being filled with a ferrite stack separated by pieces of soft iron. Results obtained with a hydrogen-filled tube under a pressure of 10^{-1} Torr. show that the system is more efficient if the magnetic field is localized near the inner electrode in this manner.

537.533.8 2910
The Theory of Secondary-Electron Emission of Metals: Part 1—G. Bimschas. (*Z. Phys.*, vol. 161, pp. 190–204; December 16, 1960.)

537.56+537.311.33 2911
Complex Conductivity of some Plasmas and Semiconductors—P. H. Fang. (*Appl. Sci. Res.*, vol. B9, no. 1, pp. 51–64; 1961.) "The complex conductivities of plasmas and semiconductors have been calculated for several cases where the collision frequency can be expressed as a power function of the energy. From the result, some characteristic parameters of the plasma originally investigated by Spitzer are estimated. The problem of determining the relaxation time from a nonsymmetrical dispersion is discussed."

537.56 2912
Couette Flow of a Fully Ionized Gas, considered as a Two-Component Fluid—L. A. Peletier and L. van Wijngaarden. (*Appl. Sci. Res.*, vol. B9, no. 2, pp. 141–150; 1961.)

537.56 2913
Experimental Study of the Dimagnetism of Gaseous Plasmas with Electron and Nuclear Spin Resonance Techniques—T. C. Marshall and L. Goldstein. (*Phys. Rev.*, vol. 122, pp. 367–376; April 15, 1961.)

537.56:538.56 2914
Energy Conversion Mechanism in a Bounded Magnetized Current-Carrying Plasma—G. H. Joshi. (*Phys. Rev. Lett.*, vol. 6, pp. 339–341; April 1, 1961.) For a finite plasma, coupling exists between quasi-transverse electromotive waves and quasi-longitudinal space-charge waves. This can lead to an energy transfer between the kinetic energy of drifting plasma and electromotive energy. This mechanism may be useful for heating a thermonuclear plasma using the "runaway electrons." Certain types of VLF radio noise can also be explained.

537.56:538.561.029.65/.66 2915
Emission of Submillimetre Electromagnetic Radiation from Hot Plasma in ZETA—G. N. Harding, M. F. Kimmitt, J. H. Ludlow, P. Porteous, A. C. Prior, and V. Roberts. (*Proc. Phys. Soc. (London)*, vol. 77, pp. 1069–1075; May 1, 1961.) The first results are reported of experiments using spectroscopy in the far infrared (0.1–1.6 mm).

537.56:538.6 2916
Oscillations of a Plasma in a Static Magnetic Field—N. Anderson. (*Proc. Phys. Soc. (London)*, vol. 77, pp. 971–979; May 1, 1961.) The dielectric tensor is calculated and used together with the solutions to the propagation problem at zero temperature to allow for thermal effects.

537.56:538.69 2917
The Theory of Plasmas in the Presence of Magnetic Fields—J. Friedrich. (*Z. Phys.*, vol. 160, pp. 494–503; November 22, 1960.) On the basis of Boltzmann's equation the complete theory of a stationary cylindrical plasma is obtained allowing for the presence of the intrinsic magnetic field and of an additional logitudinal magnetic field.

538.221 2918
Statistical Mechanics of Ferromagnetism; Spherical Model as High-Density Limit—R. Brout. (*Phys. Rev.*, vol. 122, pp. 469–474; April 15, 1961.) The conjecture of the validity of the spherical model (3789 of 1960) is examined and justified.

538.56:621.396.677.8 2919
Measurements on Resonators formed from Circular Plane and Confocal Paraboloidal Mirrors—E. H. Scheibe. (Proc. IRE, vol. 49, p. 1079; June, 1961.) A measured Q of 260,000 at 3.2 cm λ was obtained, and results were in agreement with those of Fox and Li relating to optical masers (*ibid.*, vol. 48, pp. 1904–1905: November, 1960).

538.561:537.533 2920
Cherenkov Radiation in a Dielectric Tube Waveguide—B. W. Hakki and P. D. Coleman. (Proc. IRE, vol. 49 pp. 1084–1085; June, 1961.) An experimental arrangement for measuring the radiation is described and results are given. High interaction resistances can be obtained at the expense of physical size and mode interference.

538.566:535.43 2921
Shift of the Shadow Boundary and Scattering Cross-Section of an Opaque Object—S. I. Rubinow and J. B. Keller. (*J. Appl. Phys.*, vol. 32, pp. 814–820: May 1961.)

538.566:535.43 2922
Near-Zone Back-Scattering from Large Spheres—V. H. Weston. (*Appl. Sci. Res.* vol. B9, pp. 107–116; 1961.) Backscatter from a perfectly conducting sphere for short wavelengths is theoretically treated. The back-scatter cross section is modified considerably as the receiver approaches to within a few radii, and becomes the cross section of a flat plate for a receiver very near the sphere.

538.566:539.23 2923
Reflection and Transmission of Conductive Films—M. V. Schneider. (Proc. IRE, vol. 49, pp. 1090–1091; June, 1961.) A more general relation for the reflection and transmission co-

efficients is given for the case of oblique incidence, and perturbations due to frequency dependence and displacement currents are considered. A formula is also derived for thick films. See also 4197 of 1960 (Koide).

538.569.4:537.525 2924
Microwave Zeeman Effect of Free Hydroxyl Radicals—H. E. Radford. (*Phys. Rev.* vol. 122, pp. 114–130; April 1, 1961.) The theory of the Zeeman effect in $^2\Pi$ levels of light diatomic radicals is extended to the general intermediate coupling case and used for detailed analysis of observed paramagnetic resonance absorption spectra, at 3 cm λ, of the products of an electric discharge in low-pressure H_2O and D_2O vapor.

538.569.4:538.222 2925
On the Theory of Cross-Relaxation in Maser Materials—M. Hirono. (*J. Radio Res. Labs, Japan* vol. 8, pp. 1–15; January, 1961.) The cross-relaxation rate is calculated by the method of moments, taking into account the effect of the crystalline electric field.

538.569.4:621.375.9 2926
Phonon Masers and the Phonon Bottleneck—C. Kittel. (*Phys. Rev. Lett.*, vol. 6, p. 449; May 1, 1961.) A calculation shows that the transfer of energy from the electron spin system can set up a stimulated phonon emission leading to maser action.

538.691 2927
'Corkscrew'—a Device for Changing the Magnetic Moment of Charged Particles in a Magnetic Field—R. C. Wingerson. (*Phys. Rev. Lett.*, vol. 6, pp. 446–448; May 1, 1961.) A helical field source is described which produces a monotonic increase or decrease in the transverse energy of a particle moving initially in the uniform field along the helix axis.

539.2 2928
Higher Random-Phase Approximations in the Many-Body Problem—H. Suhl and N. R. Werthamer. (*Phys. Rev.*, vol. 122, pp. 359–366; April 15, 1961.) "The usual random phase approximation combined with an equations-of-motion technique for the many-electron problem is extended yielding many of the known results of series summation methods in a straightforward manner. The method should apply to other types of many-body problems as well."

539.2:538.2 2929
A Variational Approach to Correlation in an Electron Gas—W. H. Young. (*Phil. Mag.*, vol. 6, pp. 371–377; March, 1961.) A periodic electron density is allowed, and it is proved that there exist one-particle states energetically more favorable than the usual plane-wave orbitals.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.164:621.317.794 2930
Digital Radiometer—Weinreb. (See 3115.)

523.164.3 2931
Radio Emission from the Cygnus Loop—D. S. Mathewson, M. I. Large, and C. G. T. Haslam. (*Monthly Notices Roy Astron. Soc.*, vol. 121 no. 6, pp. 543–550; 1960.) A high-resolution study of the Cygnus Loop has been made at 408 Mc and the results are compared with the optical features. Estimates of the thermal and nonthermal radio emissions are made.

523.164.3 2932
Secular Variation of the Flux Density of the Radio Source Cassiopeia A—D. S. Heeschen and B. L. Meredith. (*Nature*, vol. 190, pp. 705–706; May 20, 1961.) Observations at 1400 Mc support the evidence of Hogbom and Shakeshaft (1492 of May) of a decrease of about 1 per cent per annum in the flux density of Cassiopeia A over the last 12 years.

523.164.3:523.42 2933
The Microwave Temperature of Venus—D. E. Jones. (*Planet. Space Sci.*, vol. 5, pp. 166–167; June, 1961.) Solar corpuscular radiation may maintain an ionosphere on Venus, and the variation in observed brightness temperature may consequently be a function of zenith angle.

523.164.3:523.45 2934
The Source of Radiation from Jupiter at Decimetre Wavelengths: Part 3—Time Dependence of Cyclotron Radiation—G B. Field. (*J. Geophys. Res.*, vol. 66, pp. 1395–1405; May, 1961.) Part 2: 3471 of 1960.

523.164.4 2935
Radio Emission from the Perseus Cluster—P. R. R. Leslie and B. Elsmore. (*Observatory*, vol. 81, pp. 14–16; February, 1961.) Two small components of the Perseus cluster separated by 350 kc have been identified. They may be physically related in a similar manner to that of the two emitting regions in Cygnus-A.

523.165 2936
The Effects of Betatron Accelerat ons upon the Intensity and Energy Spectrum of Magnetically Trapped Particles—P. J. Coleman, Jr. (*J. Geophys. Res*,. vol. 66, pp. 1351–1361; May, 1961.) The effects of change of magnetic field on the energy spectrum are calculated.

523.165 2937
Characteristics and Fine Structure of the Large Cosmic-Ray Fluctuations in November 1960—J. F. Steljes H., Carmichael, and K. G. McCracken. (*J. Geophys. Res.* vo . 66, pp. 1363–1377; May, 1961.) Report and discussion of observations of solar cosmic-ray increases on November 12 and 15, the former occurring at a time of a sharp Forbush decrease.

523.165 2938
Balloon Observations of Auroral-Zone X Rays—R. R. Brown. (*J. Geophys. Res.*,vol. 66, pp. 1379–1388; May, 1961.) X rays are detectable about 10 per cent of the time; large increases occur during geomagnetic disturbances.

523.165 2939
Seasonal Variations of Cosmic-Ray Intensity in Polar Regions—K. Maeda and V. L. Patel. (*J. Geophys. Res.*, vol. 66, pp. 1389–1393; May, 1961.)

523.5:621.396.96 2940
Analysis of a Type of Rough-Trail Meteor Echo—B. A. McIntosh. (*Canad. J. Phys.*, vol. 39, pp. 437–444; March, 1961.) The characteristics of *b*-type echoes are examined in detail (see also 2226 of 1959). The rough trail may be due to fragmentation of the dust-ball type of meteoroid.

523.5:621.393.96 2941
The Height Variation of the Ambipolar Diffusion Coefficient for Meteor Trails—J. S. Greenhow and J. E. Hall. (*Planet. Space Sci.*, vol. 5, pp. 109–114; June, 1961.) Both the two-station and the split-beam measuring techniques were used. It was concluded that the atmospheric scale height during winter at a height of 98 km above Jodrell Bank (53° 14′ N) is 8.0 km

523.75:523.165 2942
Delayed Propagation of Solar Cosmic Rays on September 3, 1960—J. R. Winckler, P. D. Bhavsar, A. J. Masley, and T. C. May. (*Phys. Rev. Lett.*, vol. 6, pp. 488–491; May 1, 1961.) An analysis of balloon, rocket and ground-level observations reveals that the delay between the flare and the arrival of solar protons increases with decrease in proton energy.

523.75:523.165 2943
Rocket Observations of Solar Protons on September 3, 1960—L. R. Davis, C. E. Fichtel, D. E. Guss, and K. W. Ogilvie. (*Phys. Rev. Lett.*, vol. 6, pp. 492–494; May 1, 1961.) The results of nuclear-emulsion and G-M counter observations are used to derive the integral and differential spectra for protons of energy 10–200 Mev arriving at the top of the atmosphere.

550.385.4 2944
Magnetic Disturbances and the Earth's Magnetic Field—J. S. Chatterjee. (*J. Geophys. Res.*, vol. 66, pp. 1535–1546; May, 1961.) Storm-time changes in the earth's magnetic field could induce currents in the earth's core which interact with the steady field. This produces a nett unidirectional current at the end of the disturbance which penetrates gradually down to the core and has a decay time of a million years. This current could be responsible for maintaining the steady field.

550.385.4 2945
Solar Corpuscular Showers and Families of Geomagnetic Storms—V. I. Afanas'eva. (*Dokl. Akad. Nauk SSSR*, vol. 135, pp. 1120–1122; December 11, 1960.) Investigation of the effect of corpuscular showers in producing 1) magnetic storms and 2) magnetic disturbances during which the corpuscular showers pass near the earth without enveloping it.

550.385.4:523.165 2946
The Ring Current, Geomagnetic Disturbance, and the Van Allen Radiation Belts—S. I. Akasofu and S. Chapman. (*J. Geophys. Res.*, vol. 66, pp. 1321–1350; May, 1961.) The negative phase of a magnetic storm is ascribed to the formation of a ring current at a distance about four earth's radii from the earth's centre. The magnetic field due to the current is calculated and compared with observational results.

551.507.362.1:523.165 2947
Radiation Measurements during the Flight of the Third Space Rocket—S. N. Vernov, A. E. Chudakov, P. V. Vakulov, E. V. Gorchakov, Yu. I. Logachev, and A. G. Nikolaev. (*Dokl. Akad. Nauk SSSR*, vol. 136, pp. 322–324; January 11, 1961.) Preliminary analysis of results of instrument recordings made during the flight of the rocket between the 4th and 18th of October, 1959.

551.507.362.2 2948
Secondary-Electron Emission and the Satellite Ionization Phenomenon—D. B. Medved. (PROC. IRE, vol. 49, pp. 1077–1078; June, 1961.) Electron emission from a satellite and the electron density variation near it due to bombardment by neutral particles are considered.

551.507.362.2 2949
Sputtering as a Possible Mechanism for Increase of Ionization in the Vicinity of Low-Altitude Satellites—G. D. Magnuson and D. B. Medved. (*Planet. Space Sci.*, vol. 5, pp. 115–121; June, 1961.) An increase in ionization above ambient may occur at distances of up to 1 mean path from the satellite. Ionization potentials of the injected atoms are $\frac{1}{2}$ to $\frac{1}{5}$ those of the ambient species. The probability of any collision leading to ionization should exceed 10^{-6} with the model proposed, for the mechanism to be of interest.

551.507.362.2 2950

The Effect of Atmospheric Rotation on the Orbital Plane of a Near-Earth Satellite—G. E. Cook and R. N. A. Plimmer. (*Proc. Roy. Soc.* (*London*) *A*, vol. 258, pp. 517–528; November, 1960.) Theoretical formulas are derived for the rotation of the orbital plane and change in orbital inclination of a near-earth satellite due to the influence of the atmosphere.

551.507.362.2:523.165 2951

Satellite Determination of Heavy Primary Cosmic-Ray Spectrum—M. A. Pomerantz, S. P. Agrawal, P. Schwed and H. Hanson. (*Phys. Rev. Lett.*, vol. 6, pp. 362–364; April 1, 1961.) A preliminary report of the heavy-nuclei cosmic-ray flux observed aboard Explorer VII is given.

551.507.362.2:621.391.812.5 2952

Drop-Out Phenomenon Observed in the Satellite 1958 δ_2 Transmissions—L. Liszka. (*J. Geophys. Res.*, vol. 66, pp. 1573–1577; May, 1961.) Irradiation of the quartz oscillator in 1958 δ_2 during passage through the horns of the outer Van Allen belt may explain "drop-outs"—sudden decreases of observed signal strength.

551.507.362.2+523.3 :621.396.96 2953

Sideband Correlation of Lunar and Echo Satellite Reflection Signals in the 900-Mc/s Range—R. E. Anderson. (PROC. IRE, vol. 49, pp. 1081–1082; June, 1961.) Measurements are described and records of signal variation are illustrated. With lunar reflections a correlation of 0.1 was obtained between sidebands with 4 kc separation: the correlation of signals reflected from Echo balloon with 20-kc sideband separation was better than that of lunar reflections with 500-cps separation.

551.510.53 2954

Production of Carbon in the Upper Atmosphere—K. P. Chopra. (*Planet. Space Sci.*, vol. 5, pp. 164–166; June, 1961.) Before decaying, very fast neutrons originating from disintegrations by high energy cosmic ray particles may collide with nitrogen atoms to produce ${}_5B^{11}$ and ${}_6C^{14}$. This "neutron-capture" should result in carbon and boron atoms and traces of cyanogen gas being present in the upper atmosphere, while by-products of the mechanism may contribute significantly to the formation of the inner radiation belt.

551.510.535 2955

Studies of the *E* Layer of the Ionosphere: Part 2—Electromagnetic Perturbations and Other Anomalies—E. V. Appleton and A. J. Lyon. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 73–99; June, 1961.) Departures from the predictions of simple theory in observed *E*-layer behavior are attributed to perturbations arising from the motor effect of the S_q system of currents flowing there. The main effect is to raise f_0E near noon in low latitudes and reduce it in high latitudes. The departure of f_0E from proportionality to $(\cos \chi)^{\frac{1}{2}}$, and other anomalies are also discussed. Part 1: 1432 of 1957.

551.510.535 2956

Note on the Role of Small Layers with Irregular Surfaces in the Formation of Sporadic-E Layers—J. Voge. (*Ann. Télécommun.*, vol. 16, pp. 62–63; January/February, 1961.) In earlier work by du Castel *et al.* (*ibid.*, vol. 15, pp. 38–47 and 107–142; January/February and May/June, 1960) irregularities in the refractive index of the troposphere have been represented by a model which has now been adapted to the E_s layer. The reflection coefficients calculated using the model are of the right order of magnitude.

551.510.535 2957

Irregularities in the E_s Layer of the Ionosphere—G. F. Fooks. (*Nature*, vol. 190, p. 707–708; May 20, 1961.) Correlation analysis shows no systematic elongation at Cambridge (52°N, 0°E) in contrast to the E-W elongation of auroral, and N-S elongation of equatorial type of E_s. There is a wide range of sizes in the scale of the irregularities in the diffraction pattern on the ground.

551.510.535 2958

Consideration of the F_1 Layer in the Ionosphere—S. Ishikawa and S. Hiidome. (*J. Radio Res. Labs., Japan*, vol. 8, pp. 29–40; January, 1961.) A graphical statistical description of the diurnal, seasonal and annual variations of f_0F_1 at Kokobunji and information on the relation between f_0F_1 and f_0F_2.

551.510.535 2959

Diurnal Variation of F_1-Region Drifts at Waltair—E. B. Rao and B. R. Rao (*Curr. Sci.*, vol. 30, pp. 9–10; January, 1961.) The results of F_1-region drift measurements for the four seasons in the years 1956–1958 are given. See also 2242 of 1959.

551.510.535 2960

Equatorial Spread-F at Ibadan, Nigeria—A. J. Lyon, N. J. Skinner, and R. W. Wright. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 100–119; June, 1961.) Spread-F characteristics are closely related to the height of the layer; the irregularities may be due to hydromagnetic disturbances, which are increasingly attenuated at lower heights. See also 879 of March.

551.510.535 2961

Variation in Height of Anisotropy and Random Drift Velocity of the Irregularities in the Ionosphere—B. R. Rao and K. V. V. Ramana. (*Nature*, vol. 190, no. 4777, pp. 706–707; May, 1961.) Correlation analysis shows that the axial ratio of the anisotropy ellipse and the random change factor $(V_C)_v/V$ are both greatest for waves reflected from heights of 270–290 km.

551.510.535 2962

Possible Identification of Atmospheric Waves Associated with Ionospheric Storms—M. D. Wright. (*Nature*, vol. 190, no. 4779, pp. 898–899; June, 1961.) The onset time of an ionospheric storm is defined as the time of first increase at $h'F$ and the velocity is determined by measuring the time delay at Godley Head, 1105 km north of Campbell Island. The mean northward velocity for eight well-defined storms was found to be in the range 406–691 m/sec and may correspond to the velocity of a neutral gaseous wave as indicated by the Johnston Island nuclear explosion disturbance, *i.e.*, 470 m/sec.

551.510.535 2963

A Method of Calculation of N(h) Profiles from Ionospheric $h'(f)$ Curves—H. Hojo. (*J. Radio Res. Labs., Japan*, vol. 8, pp. 41–57; January, 1961.) See 1509 of May.

551.510.535 2964

A New Method for the Analysis of Ionospheric $h'(f)$ Records—J. E. Titheridge. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 1–12, April, 1961.) A method is given for calculating the real heights of reflection in the ionosphere, from the observed virtual heights. The real height is assumed to be a polynomial in the plasm, frequency.

551.510.535 2965

General Expression for the Group Refractive Index of the Ionosphere—V. S. N. Murty and S. R. Khastgir. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 65–69; April, 1961.) A general expression for the group refractive index of the ionosphere is found in which wave frequency, electron density and collision frequency, magnetic field and direction of propagation are variables. See also 2001 of 1960.

551.510.535:517.64 2966

A Solution of the Integral Equation $h'(f) = \int \mu'(f; f_0) dz(f_0)$—H. Unz. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 40–45; April, 1961.) A method involving power series expansions is given for solving the integral equation. One advantage of the method is the use of a set of master curves which applies for all stations.

551.510.535:523.745 2967

A Study of the Variations in a Daily Ionospheric Index of Solar Activity—G. H. Bazzard. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 193–202; June, 1961.) A table of daily values of $J_E f_E^4 \sec \chi$ at Slough is given for 1959. The short term variations in J_E from July, 1957–December, 1959 are examined by statistical methods and it is shown that these variations are related to changes in the sun.

551.510.535:523.746 2968

A Long-Term Variation in the Relationship of Sunspot Numbers to *E*-Region Character Figures—R. Naismith, H. C. Bevan and P. A. Smith. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 167–173; June, 1961.) The statistical relation between sunspot number and month mean values of $Ch_E = f_E^4 \sec \chi$ at Slough suggests that from 1933 onwards, there has been a decrease in the value of Ch_E corresponding to a given sunspot number.

551.510.535:523.75 2969

Solar Flare Effects in the *F* Region of the Ionosphere—R. W. Knecht and K. Davies. (*Nature*, vol. 190, no. 4778, pp. 797–798; May, 1961.) Increases in f_0F_2 were observed during flares in 1949 and 1956 which were accomplnied by increases in sea-level cosmic-ray intensity. Similar increases in f_0F_2 are reported in November, 1960, at two stations and simultaneous falls in isoionic heights of 10–20 km were also observed.

551.510.535:539.16 2970

Sporadic-*E* Phenomena associated with the High-Altitude Nuclear Explosions over Johnston Island—L. Thomas and R. E. Taylor. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 205–208; June, 1961.) Marked increases in E_s occurred near latitudes 60° N and S and are attributed to the explosions.

551.510.535:550.385.2 2971

Magnetic Effects in the *F* Region of the Ionosphere—J. W. King (*J. Atmos. Terr. Phys.*, vol. 21, pp. 26–34; April, 1961.) Nighttime values of f_0F_2 at stations in the south of Africa are found to decrease with increasing values of the equivalent planetary daily amplitude A_p. The magnetic control of nighttime ionization could account for the day-to-day correlation of f_0F_2 values. The relation of ionospherically and magnetically quiet days is discussed.

551.510.535:621.3.085 2972

Automatic Ionogram Scaler—M. Onoye, F. Ochi, N. Omiya and K. Arai. (*J. Radio Res. Labs, Japan*, vol. 8, pp. 59–63; January, 1961.) An instrument has been designed which projects an ionogram onto a screen. The parameters which are read off are recorded simultaneously on punched tapes and in typewritten form. The tape can be used to produce punched cards if required.

551.510.535:621.3.087.4 2973

Active High-Frequency Spectrometers for Ionospheric Sounding: Part 3—Determination

of True Height from the Virtual Height of Reflection—A. K. Paul. (*Arch. elekt. Übertragung*, vol. 14, pp. 468–476; October, 1960.) A new method for determining true height is described (see also 2591 of August) and its application to daytime and nighttime soundings discussed. Part 2: 2597 of August (Rawer).

551.510.535:621.3.087.4 2974
A New Type of Ionospheric Drift Recorder.—B. R. Rao and R. R. Rao. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 208–210; June, 1961.) A method of eliminating photographic and pen recording of fading signals and the laborious calculations. A tape recorder is used to introduce a variable time delay into one antenna feed. When this delay equals that of the received signals a minimum is obtained. Delays of 0.07–1.5 sec have been measured.

551.510.535:621.391.812.63 2975
Day-to-Day and Station-to-Station Correlation of Ionospheric F-Region Critical Frequencies—J. W. King. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 35–39; April, 1961.) The correlation of simultaneous f_0F_2 values at stations separated by 1700 km in a north-south direction is found to have a semi-diurnal variation, the phase of which varies with season. Results at a single station show an important autocorrelation for values at the same time on consecutive days, and this is ascribed to magnetic effects.

551.510.535:621.391.812.63 2976
Year-to-Year Changes in the Winter Anomaly in Ionospheric Absorption—L. Thomas. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 215–217; June, 1961.) The mean anomalous component in a given winter is linearly related to the mean magnetic C figure for the same winter.

551.510.535(98):523.75 2977
Some Characteristics of Solar Corpuscular Radiations which Excite Abnormal Ionization in the Polar Upper Atmosphere—K. Sinno. (*J. Radio Res. Labs. Japan*, vol. 8, pp. 17–28; January, 1961.) Statistical data on the time delay from the occurrence of a major flare until the onset of polar absorption have been studied. It is suggested that particles ejected from the sun may be trapped in a corpuscular cloud which may later intersect a magnetic line of force along which particles can travel to reach the earth.

551.510.536 2978
The Coupling between the Protonosphere and the Normal F Region—W. B. Hanson and I. B. Ortenburger. (*J. Geophys. Res.*, vol. 66, pp. 1425–1435; May 1961.) In the protonosphere, above about 900 km, the distribution of protons is determined by diffusion rather than chemical processes: this region is independent of diurnal changes in the ionosphere.

551.510.62 2979
Continuous Recordings for the Determination of the Vertical Gradient of the Refractive Index in the Free Atmosphere—H. P. Barthelt and A. Rossbach. (*Nachrtech. Z.*, vol. 13, pp. 462–464; October, 1960.) Report on measurements made by equipment mounted at six different levels on a 160-m tower to determine the correlation between the refractive-index gradient and field-strength fluctuations over a 250-km radio-link path.

551.510.62 2980
The Radio Refractive Index Gradient over the British Isles—J. A. Lane. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 157–166; June, 1961.) Airborne microwave refractometer measurements made in the British Isles confirm that the quantity $(n-1)$ varies exponentially with height. The variations in the parameters are less than those in the U. S. There is a high correlation between the surface value of $(n-1)$ and the average decrease in the refractive index in the first kilometre in altitude.

551.594.21 2981
A Criterion for Thunderstorm Theories—J. A. Chalmers. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 174–176; June, 1961.) Results on the charging and external currents of the same storm favor the classical theory rather than the convection theories. See 165 of 1960.

551.594.5 2982
Aurora and Airglow from Colour Film Observations—B. P. Sandford. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 177–181; June 1961.) Summary of experience using color film records.

551.594.5:621.396.96 2983
Investigation of Auroral Echoes: Part 2.—L. Harang and J. Tröim. (*Planet Space Sci.*, vol. 5, pp. 105–108; June, 1961.) Ratios of measured echo signal strength at 40 and 80, and 40 and 120 Mc showed that the frequency dependence of the reflection coefficient is qualitatively explained by either an inverse λ^2 law (partial reflection from large surfaces) or an exp $(-k/\lambda^2)$ law (scattering from field-aligned irregularities) Part 1: 1175 of April.

551.594.6 2984
Narrow-Band Atmospherics from Two Local Thunderstorms—F. Horner. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 13–25; April, 1961.) Atmospherics from two storms were studied. Reception frequencies of 11 Mc and 6 kc were used with power bandwidths in the range 200–300 cps. Duration, peak amplitude, mean field strength and mean power flux were calculated. Results conflict with some theories of the origin of atmospherics.

551.594.6 2985
On the Extremely-Low-Frequency Spectrum of Earth-Ionosphere Cavity Response to Electrical Storms—H. R. Raemer. (*J. Geophys. Res.*, vol. 66, pp. 1580-1583; May, 1961.) A comparison of the observed and calculated spectra in the range 5 -33 cps.

551.594.6 2986
Simultaneous Observations of V.L.F. Noise ('Hiss') at Hobart and Macquarie Island—R. L. Dowden. (*J. Geophys., Res.*, vol. 66, pp. 1587–1588; May, 1961.) The noise sources seem to be small in size and to have different locations at 4 and 9 kc.

551.594.6 2987
Dispersions of Multiple-Stroke Whistlers—N. D. Clarence and P. A. O'Brien. (*Nature*, vol. 190, pp. 1096–1097; June 17, 1961.) Analysis of I.G.Y. records at Durban indicates a 4–6 per cent increase in the dispersion of successive components of multiple-stroke short whistlers which supports the current-jet hypothesis of whistler generation.

LOCATION AND AIDS TO NAVIGATION

621.396.663.089.6 2988
Problems of Calibration of Two-Channel Cathode-Ray Direction Finders—G. Kattner and W. Rohrbeck. (*Nachrtech.*, vol. 10, pp. 392–397; September, 1960.) Methods of calibrating and checking the calibration of direction finders are outlined, including one for periodically interchanging the two channels.

621 396.932.2:627.95 2989
Marine Distress Beacon—K. Jirsa. (*Telefunken Ztg.*, vol. 33, pp. 290–294; December, 1960.) Description of a floating distress beacon operating at 2.182 Mc using battery-driven transistor circuits and the ferrite-rod antenna detailed in 4223 of 1959 (Baur and Ziehm).

621.396.963+534.88 2990
Video Integration in Radar and Sonar Systems—D. C. Cooper and J. W. R. Griffiths. (*J. Brit. IRE*, vol. 21, pp. 421–433; May, 1961.) The performance of ideal and practical systems, including the influence of antenna beamwidth, is examined. A new double delay-loop integrator giving a small improvement in threshold detection and having some practical advantages is described.

621.396.963.325 2991
Nonlinear Effects in Rotating-Coil P.P.I. Displays—R. H. C. Morgan . (*Mullard Tech. Commun.*, vol. 5, pp. 252–258; March, 1961.) "Nonlinearity effects of the three coaxial deflection coils in a PPI display system and methods to minimize them by means of careful coil design and adjustment, are discussed."

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 2992
Theory of the Photomagnetic effect in Cubic Anisotropic Crystals—A. A. Grinberg. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1361–1367; July, 1960.)

535.215:537.311.33 2993
Determination of Recombination Contents from the Spectral Characteristics of a Photocell with a *p-n* Junction: Part 2—G. B. Dubrovskiĭ and V. K. Subashiev. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1562–1571; July, 1960.) The analysis of Part 1 [1871 of June (Subashiev)] is applied to photocells obtained by diffusion of donor impurities in p-type Si.

535.215:539.23 2994
Photoconductivity of Sulphur Layers Treated with Mercury Vapour—M. I. Korsunskiĭ, N. S. Pastushuk, and G. D. Mokhov. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1581–1583; July, 1960.) Treated layers of resistance $\lessdot 10^6\ \Omega$, showed simultaneous positive and negative photoconductivity of large inertia. In the short-wavelength region ($\lambda = 453\ m\mu$) negative photoconductivity predominated; in the long-wavelength region ($\lambda = 645\ m\mu$) only positive photoconductivity was observed.

535.215:546.47'221 2995
An Investigation of the Induced Conductivity in Thin Zinc Sulphide Layers—A. A. Didenko, Yu. A. Nemilov, and V. I. Fomina. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1434–1440; July, 1960.)

535.215:546.48'221 2996
The Contribution of Ions to the Current through CdS at High Temperature—K. W. Böer and H. Gutjahr. (*Mber. Dtsch. Akad. Wiss. Berlin*, vol. 2, no. 6, pp. 324–327; 1960.) Experimental investigations indicate that up to a temperature of 430°C, the ion current is less than 0.1 per cent. See also 2241 of July (Böer, *et al.*).

535.215:546.48'221 2997
An Investigation of the Lux/Ampere Characteristics of CdS Single Crystals—E. A. Sal'kov and G. A. Fedorus. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1576–1580; July, 1960.) Results of photocurrent measurements are at variance with theory [see, *e.g.*, 768 of 1956 (Rose) and 3102 of 1958 (Bube)].

535.215:546.48'221 2998
Intermittent-Light Method for the Measurement of Capture Cross-Sections for Traps in Cadmium Sulphide—E. A. Niekisch. (*Z.*

Phys., vol. 161, pp. 38–45; December 1, 1960.) The method described in 2972 of 1955 is used to determine capture cross-sections for shallow traps near the conduction band in CdS crystals.

535.215.546.48'221 2999
Current Noise and Distributed Traps in Cadmium Sulphide—J. J. Brophy. (*Phys. Rev.*, vol. 122, pp. 26–30; April 1, 1961.) Measurements of noise as a function of frequency in a single crystal, at different temperatures and levels of illumination, give a $1/f$ trapping noise spectrum of the form $(1/\omega)$ arc tan $\omega\tau$. Values of τ are obtained from the low-frequency turnover point, and give trap depths which are in good agreement with discrete trapping levels determined from other measurements.

535.215:546.48'221 3000
Fine Structure and Magneto-optic Effects in the Exciton Spectrum of Cadmium Sulphide—J. J. Hopfield and D. G. Thomas. (*Phys. Rev.* vol. 122, pp. 35–52; April 1, 1961.) Detailed experimental study of the magneto-optic absorption spectrum of direct excitons formed from the top valence band and the conduction band, and theoretical interpretation of results. See also 1641 of 1960 (Thomas and Hopfield).

535.215:546.48'241 3001
The Optical Properties of Cadmium Telluride in the Far Infrared Region—A. Mitsuishi. (*J. Phys. Soc., Japan*, vol. 16, pp. 533–537; March, 1961.)

535.37:546.47'221:535.215 3002
Impedance Changes Induced by Light in ZnS Phosphors—I. Uchida and K. Satake. (*J. Phys. Soc. Japan*, vol. 16, p. 573; March, 1961.) The impedance change of a capacitor containing a phosphor is interpreted in terms of an equivalent circuit and used to derive the photoconductivity.

535.376 3003
A New Electroluminescence Effect in Black Carborundum—A. G. Gol'dman. (*Dokl. Akad. Nauk SSSR*, vol. 135, no. 5, pp. 1108–1110; December, 1960.) Samples are prepared by applying the p-type material a sensitive n-type layer. When acted upon by current in the forward direction a luminescence is produced at the p-n junction.

535.376 3004
The Field Enhancement of ZnS.CdS-Mn Phosphors with X-Ray Excitation—H. Winkler, H. Rĕppischer, and G. Wendel. (*Z. Phys.*, vol. 161, no. 3, pp. 330–338; December, 1960.) The results of experimental investigations differ to some extent from those of other authors. The "memory" effect described by Destriau (1213 of 1959) was observed as a consequence of field action alone before X-ray excitation.

535.376:546.47'221 3005
Direct-Current Electroluminescence at Low Voltages—W. A. Thornton. (*Phys. Rev.*, vol. 122, pp. 58–59; April 1, 1961.) Electroluminescence can occur at applied voltages corresponding to only about half the band gap of a ZnS phosphor; hence the acceleration-collision theory is ruled out at low voltage.

535.376:546.47–31 3006
Excitation of Zinc Oxide by Slow Electrons—P. Wachter. (*Z. Phys.*, vol. 161, pp. 62–73; December 1, 1960.) Investigations of low-energy excitation of ZnO phosphors were carried out with an electron beam of small velocity spread; a theoretical interpretation of the results is given.

537.226:546.47'221:539.23 3007
On the Dielectric Properties of Thin Films of Zinc Sulphide—R. Fuchshuber, R. Gullien, and S. Roizen. (*C.R. Acad. Sci., Paris*, vol. 251, pp. 51–53; July 4, 1960.)

537.227 3008
Dielectric Properties of Solid Solutions of Sodium Niobate-Lead Zirconate and Sodium Niobate-Lead Titanate—W. R. Bratschun and R. L. Cook. (*J. Am. Ceram. Soc.*, vol. 44, no. 3, pp. 136–140; March, 1961.)

537.227 3009
Variation of Permittivity with Electric Field in Perovskite-Like Ferroelectrics—H. Diamond. (*J. Appl. Phys.*, vol. 32, pp. 909–915; May, 1961.) Variation of incremental permittivity is shown to be associated with an induced ferroelectric state rather than being directly a property of domain processes; large variations with field must be accompanied by strong thermal sensitivity.

537.227 3010
Bismuth Titanate: a Ferroelectric—L. G. Van Uitert and L. Egerton. (*J. Appl. Phys.* vol. 32, p. 959; May, 1961.) Results of measurements of dc resistivity, dielectric constant, and hysteresis loops are given.

537.227:546.431'824–31 3011
Optical Absorption and Electrical Conductivity of Reduced $BaTiO_3$ Single Crystal—S. Ikegami, I. Ueda, and Y. Ise. (*J. Phys. Soc., Japan*, vol. 16, pp. 572–573; March, 1961.)

537.227:547.476.3 3012
The Effect of a Constant Electric Field on the Hysteresis of Rochelle Salt—K. N. Karmen. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1671–1675; July, 1960.) A circuit used for recording hysteresis curves during simultaneous application of steady and variable fields of different intensities is shown. Symmetric and asymmetric hysteresis loops are identified and discussed.

537.227:547.476.3:539.12.04 3013
Ferroelectric Properties of X-Ray-Damaged Rochelle Salt—K. Okada. (*J. Phys. Soc., Japan*, vol. 16, pp. 414–423; March, 1961.)

537.311.33 3014
Semiconducting Properties of the $AgFeTe_2$ Phase—E. L. Shtrum. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1489–1493; July, 1960.)

537.311.33 3015
The Effect of Trapping Levels in Semiconductors on Steady-State Photoconductivity and on the Life time of Nonequilibrium Current Carriers—A. A. Grinberg, L. G. Paritskiĭ, and S. M. Ryvkin. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1545–1561; July, 1960.) A quantitative investigation is made of the effects of trapping levels introduced into the crystal on carrier recombination under steady-state conditions due to the presence of other traps in the forbidden zone. The effect of the filling of trapping levels on the variation of carrier lifetimes and photoconductivity as a function of temperature and light intensity is discussed.

537.311.33 3016
Long-Range Interactions between Magnetic Moments in Semiconductors—W. Baltensperger and A. M. de Graaf. (*Helv. Phys. Acta*, vol. 33, no. 8, pp. 881–888; November, 1960.) The interaction between magnetic moments in a nondegenerate electron gas is calculated, and account is taken of the additional interaction arising from virtual excitations of the valence band in semiconductors.

537.311.33 3017
Micro-masking for Chemical Etching of Semiconductors—S. Denda. (*Solid-State Electronics*, vol. 2, pp. 69; January 1961.) A method of producing a picein mask down to 30-μ diameter.

537.311.33 3018
Theoretical Calculations of the Enthalpies and Entropies of Diffusion and Vacancy Formation in Semiconductors—R. A. Swalin. (*J. Phys. Chem. Solids*, vol. 18, pp. 290–296; March, 1961.) The covalent-bond approach is used to obtain kinetic parameters.

537.311.33 3019
One-Dimensional Overlap Functions and their Application to Auger Recombination in Semiconductors—A. R. Beattie and P. T. Landsberg. (*Proc. Roy. Soc.* (*London*), vol. 258, pp. 486–495; November 8, 1960.) Certain overlap integrals, occurring as a temperature-independent parameter in a previous paper (3341 of 1959), are evaluated. Used in conjunction with the previous theory, good agreement is obtained with experimental results.

537.311.33 3020
Theory of the Auger Neutralization of Ions at the Surface of a Diamond-Type Semiconductor—H. D. Hagstrum. (*Phys. Rev.*, vol. 122, pp. 83–113; April 1, 1961.)

537.311.33 3021
Semiconducting Properties of Inorganic Amorphous Materials—H. L. Uphoff and J. H. Healy. (*J. Appl. Phys.*, vol. 32, pp. 950–954; May, 1961.) Nine materials in the systems As:Se:Te and As:S:Te were tested. Resistivity varied exponentially with temperature; the Seebeck coefficient varied linearly.

537.311.33:535.211 3022
Photothermal Effect in Semiconductors—W. W. Gärtner. (*Phys. Rev.*, vol. 122, pp. 419–424; April 15, 1961.) The establishment of a temperature distribution in a solid by optically excited diffusing and recombining carriers is called the photothermal effect. Theory is developed.

537.311.33:538.567 3023
Detection of Millimetre- and Submillimetre-Wave Radiation by Free-Carrier Absorption in a Semiconductor—B. V. Rollin. (*Proc. Phys. Soc.*, vol. 77, pp. 1102–1103; May 1, 1961.) The change in conductivity when radiation is absorbed by free carriers can be used for detection at mm and sub-mm λ. The sensitivity for n-type InSb at a temperature of about 2°K is estimated.

537.311.33:538.614 3024
Theory of Interband Faraday Rotation in Semiconductors—B. Lax and Y. Nishina. (*Phys. Rev. Lett.*, vol. 6, pp. 464–467; May 1, 1961.) The interband Faraday rotation and its dependence on frequency is examined by using a quantum-mechanics approach to calculate the rotations for the indirect and direct forbidden transitions.

537.311.33:538.63 3025
Influence of Conductivity Gradients on Galvanomagnetic Effects in Semiconductors—R. T. Bate and A. C. Beer. (*J. Appl. Phys.*, vol. 32, pp. 800–805; May, 1961.) An approximate solution is found of a boundary-value problem arising from the continuity equation in an inhomogeneous semiconductor. Experimental results indicate that transverse currents, which do not occur in the simple case discussed, do appear in general.

537.311.33:546.24 3026
Infrared Absorption and Structure of the Hole Band of Tellurium—V. M. Korsunskiĭ and M. P. Lisitsa. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1619–1623; July, 1960.)

537.311.33:546.28 3027
Absorption of Infrared Radiation by Electrons in p-Type Silicon—V. A. Yakovlev. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1624–1628; July, 1960.)

537.311.33:546.28 3028
Problems related to p-n Junctions in Silicon—W. Shockley. (*Solid-State Electronics*, vol. 2, pp. 35–67; January, 1961.) A simple model of the behavior of holes and electrons in the high reverse fields of p-n junctions is shown to account well for secondary ionization, avalanche breakdown and microplasma phenomena. Statistical spatial fluctuations of donors and acceptors are shown to have significant effects on avalanche breakdown. Typical noisy microplasma phenomena are probably associated with structural defects whose properties are specified.

537.311.33:546.28 3029
Impurity Effect upon Mobility in Heavily Doped Silicon—Y. Furukawa. (*J. Phys. Soc., Japan*, vol. 16, p. 577; March 1961.) A similar effect was obtained to that previously observed in heavily doped Ge (see 3941 of 1960).

537.311.33:546.28 3030
Some Experiments using a Vacuum-Cleaned Silicon p-n Junction—J. T. Law. (*J. Appl. Phys.*, vol. 32, pp. 848–855; May, 1961.) Changes in the junction characteristics during the adsorption of oxygen and hydrogen have been investigated. In the clean condition the valence band edge was 0.13–0.14 ev below the Fermi level. A large excess current across the clean junction disappeared during absorption of gas.

537.311.33:546.28:539.12.04 3031
The Effect of Fast-Electron Bombardment on the Electrical Conductivity of Silicon and the Dependence of the Rate of Defect Formation on the Orientation of a Crystal with respect to the Electron Beam—V. S. Vavilov, V. M. Patskevich, B. Ya. Yurkov, and P. Ya. Glazunov. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1431–1433; July, 1960.)

537.311.33:546.289 3032
The Mechanism of Carrier Scattering in p-Type Germanium—M. N. Vinogradova, O. A. Golikova, B. P. Mitrenin, and L. S. Stil'bans. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1428–1430; July, 1960.)

537.311.33:546.289 3033
Effect of Linear Dislocations on Charge Carrier Recombination in p-Type Germanium—M. I. Iglitsyn and L. I. Kolesnik. (*Fiz. Tverdogo Tela.*, vol. 2, pp. 1542–1544; July, 1960.)

537.311.33:546.289 3034
Investigation of the Kinetics of Fast Surface States on Germanium—V. G. Litovchenko and V. I. Lyashenko. (*Fiz. Tverdogo Tela.*, vol. 2. pp. 1592–1596; July, 1960.) Report of field-effect and photoconductivity measurements on high resistivity n- and p-type Ge.

537.311.33:546.289 3035
Measurements of the Anomalous Transmission Factor and of the Reflection of X Rays on Good Germanium Single Crystals for the Bragg Case of Interference: Comparison with the Dynamic Theory of X-Ray Diffraction—U. Bonse. (*Z. Phys.*, vol. 161, pp. 310–329; December 28, 1960.)

537.311.33:546.289 3036
Impurity Concentrations in the Regrown Region of In-Ge Alloyed Junctions—F. E. Roberts. (*Solid-State Electronics*, vol. 2, pp. 8–13; January, 1961.) The development of an infrared microscope for routine carrier-density measurements on small semiconducting specimens is described. The results of a study made with it of In-Ge and In-Ga-Ge recrystallized alloyed regions are given.

537.311.33:546.289 3037
Regrowth of Germanium Single Crystal from Indium Melt—M. Tomono. (*J. Phys. Soc., Japan*, vol. 16, pp. 439–453; March, 1961.)

537.311.33:546.289 3038
Hot-Electron Nonequilibrium Carrier Distribution in a Many-Valley Semiconductor—L. Gold. (*J. Phys. Soc., Japan*, vol. 16, pp. 575–576; March, 1961.) Results of a study of the contribution of nonuniform charge density to the ellipsoidal energy surfaces in Ge are given. A reasonable deviation of collinear current and voltage directions is obtained; thus measurements of this quantity may be useful in studies of effective mass and carrier density.

537.311.33:546.289 3039
Cyclotron Resonance in Germanium—R. R. Goodman. (*Phys. Rev.*, vol. 122, pp. 397–405; April 15, 1961.) A comparison is made of cyclotron resonance theory and experimental data [see 1093 of 1956 (Fletcher *et al.*)]. Values of effective mass constants which best fit the data are found

537.311.33:546.289 3040
Generation—Recombination Noise in p-type Gold-Doped Germanium—L. J. Neuringer and W. Barnard. (*Phys. Rev. Lett.*, vol. 6, pp. 455–457; May 1, 1961.) A theory is developed which accounts for the noise observed in interacting two-level systems and shows good quantitative agreement with experimental observations on Au-doped Ge.

537.311.33:546.289 3041
Third-Order Elastic Moduli of Germanium—T. Bateman, W. P. Mason, and H. J. McSkimin. (*J. Appl. Phys.*, vol. 32, pp. 928–936; May, 1961.)

537.311.33:546.289:535.215 3042
A New Effect of Negative Photoconductivity in a Magnetic Field—A. A. Grinberg, S. R. Novikov, and S. M. Ryvkin. (*Dokl. Akad. Nauk SSSR*, vol. 136, pp. 329–331; January 11, 1961.) A note on effects observed in a p-type Ge sample illuminated in the direction of an applied magnetic field.

537.311.33:546.289:535.215 3043
Modulation of Light by means of an Electric Field—B. H. Claussen. (*Proc. Phys. Soc.*, vol. 77, pp. 1100–1101; May 1, 1961.) Experiments are noted on the modulation of light reflected at a Ge surface when the carrier density was varied by means of an applied electric field. Results show that the modulation was a function of the dielectric used as a field plate rather than a function of the Ge surface.

537.311.33:546.289:538.614 3044
Faraday Effect for Direct Magneto-optical Transitions in Germanium—M. Suffczynski. (*Proc. Phys. Soc.*, vol. 77, pp. 1042–1045; May 1, 1961.) The Faraday effect due to the interband magneto-optical transitions in Ge is calculated. The direct transitons from the upper-most Landau levels of the two light-hole ladders and two heavy-hole adders are considered. Measurement of the Faraday rotation offers a sensitive means of studying the effective masses of the levels in the presence of the magnetic field.

537.311.33:546.289:538.63 3045
Galvanomagnetic Effects in n-Ge in the Impurity Conduction Range—R. J. Sladek and R. W. Keyes. (*Phys. Rev.*, vol. 122, pp. 437–442; April 15, 1961.) The measured magnitude and the crystalline anisotropy of the magnetoresistance are interpreted in terms of the changes in the donor wave functions produced by the magnetic field. The field dependence of the Hall coefficient is interpreted as a magnetoresistance effect of the conduction band.

537.311.33:546.289:539.12.04 3046
The Action of Radiation on Semiconductors: Recombination Centres Introduced in Germanium by 2-MeV Electrons—P. Baruch. (*Ann. Phys.*, (*Paris*), vol. 6, pp. 21–79; January /February, 1961.)

537.311.33:546.289:539.23 3047
Textural Properties of Germanium Films—J. E. Davey. (*J. Appl. Phys.*, vol. 32, pp. 877–880; May, 1961.)

537.311.33:546.26–1 3048
Some Optical and Electrical Properties of Semiconducting Diamonds—A. Halperin and J. Nahum. (*J. Phys. Chem. Solids*, vol. 18, pp. 297–306; March, 1961.)

537.311.33:546.3–1'289'28 3049
Magnetic Susceptibility of Ge-Si Mixed Crystals—G. Busch and O. Vogt. (*Helv. Phys. Acta*, vol. 33, no. 8, pp. 889–910; November, 1960. In German.) Susceptibility was measured as a function of temperature for 34 samples in the temperature range 300–1200°K. The results are compared with those of measurements of cyclotron resonance and magnetoresistance.

537.311.33:546.3–1'289'28 3050
Lattice Thermal Conductivity of Germanium-Silicon Alloy Single Crystals at Low Temperatures—A. M. Toxen. (*Phys. Rev.*, vol 122, pp. 450–458; April 15, 1961.

537.311.33:546.431'824–31:621.316.825.2 3051
Semiconducting Bodies in the Family of Barium Titanates—O. Saburi. (*J. Amer. Ceram. Soc.*, vol. 44, pp. 54–63; February 1, 1961.) The resistivity/temperature characteristics of the semiconducting Ba titanates are examined to determine compositons suitable for the construction of positive-type thermistors.

537.311.33:546.47'241:546.48'221 3052
Some Electrical Properties of Zinc Telluride-Cadmium Sulphide Heterojunctions—M. Aven and D. M. Cook. (*J. Appl. Phys.*, vol. 32, pp. 960–961; May, 1961.) Techniques for preparing the junctions at room temperatures are described. Rectification ratios up to 5×10^4 were obtained with a CdS film and about 10^2 with a ZnTe film. The I/V characteristic of a CdS-film junction is given.

537.311.33:546.48'19 3053
Galvanomagnetic Properties of n-Type $CdAs_2$—A. S. Fischler. (*Phys. Rev.*, vol. 122, pp. 425–429; April 15, 1961.)

537.311.33:546.561–31 3054
Electrical Conductivity of Single-Crystal Cuprous Oxide at High Temperatures—R. S. Toth, R. Kilkson, and D. Trivich. (*Phys. Rev.*, vol. 122, pp. 482–488; April 15, 1961.)

537.311.33:546.681'86 3055
Optical Investigation of the Band Structure of GaSb—M. Cardona. (*Z. Phys.*, vol. 161, no. 1, pp. 99–102; December, 1960.) The reflectivity of GaSb was measured at photon energies between 0.5 and 5 ev; the structure observed is similar to that of Ge.

537.311.33:546.681'86 3056
Distribution Coefficients of Impurities in Gallium Antimonide—R. N. Hall and J. H.

Racette. (*J. Appl. Phys.*, vol. 32, p. 856; May, 1961.)

537.311.33:546.682'18 3057
Optical Studies of the Band Structure of InP—M. Cardona. (*J. Appl. Phys.*, vol. 32, p. 958; May, 1961.)

537.311.33:546.682'86 3058
Infrared Cyclotron Resonance in InSb—E. D. Palik, G. S. Picus, S. Teitler, and R. F. Wallis. (*Phys. Rev.*, vol. 122, pp. 475–481; April 15, 1961.)

537.311.33:546.682'86:538.569.4 3059
Cyclotron Resonance in Indium Antimonide at High Magnetic Fields—B. Lax, J. G. Mavroides, H. J. Zeiger, and R. J. Keyes. (*Phys. Rev.*, vol. 122, pp. 31–35; April 1, 1961.) An improved theoretical treatment is given which accounts satisfactorily for various sets of experimental data.

537.311.33:546.682'86:538.63 3060
Magnetoresistance and Hall Effect in Oriented Single-Crystal Samples of *n*-Type Indium Antimonide—C. H. Champness. (*Canad. J. Phys.*, vol. 39, pp. 452–467; March, 1961.)

537.311.33:546.682'86:538.63 3061
Influence of Magnetoconductivity Discontinuities on Galvanomagnetic Effects in Indium Antimonide—R. T. Bate, J. C. Bell, and A. C. Beer. (*J. Appl. Phys.*, vol. 32, pp. 806–814; May, 1961.) Observations are predicted qualitatitvely by considering a simple model consisting of a long thin specimen having a discontinuity in resistivity and Hall coefficient in the current direction. The boundary-value problem is solved for this case.

537.311.33:546.683'231 3062
Certain Properties of Thallium Selenide Single Crystals—G. A. Akhundov, G. B. Abdullaev, and G. D. Guseĭnov. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1518–1521; July, 1960.)

537.311.33:546.873'221 3063
Some Semiconducting Properties of Bismuth Trisulphide—L. Gildart, J. M. Kline, and D. M. Mattox. (*J. Phys. Chem. Solids*, vol. 18 pp. 286–289; March 1961.) Preparation of single crystals dendrites and crystalline films is described. Data are given for the thermal energy gap of films optical energy gap resistivity carrier concentration and mobility, thermal conductivity and Seebeck coefficient for single crystals.

537.311.33:621.391.822 3064
Approximate Solution to Semiconductor Noise as a Queuing Problem—T. L. Saaty. (Proc. IRE vol. 49 p. 1095; June 1961.) An alternative treatment to that of Bell (3897 of 1958) indicating the possible usefulness of the queuing representation for low frequencies.

537.312.62 3065
The Surface Impedance of Normal and Superconducting Indium at 3000 Mc/s—P. N. Dheer. [*Proc. Roy. Soc.* (*London*) *A*, vol. 260, no. 1302 pp. 333–349; March 1961.] From impedance measurements, a value is obtained for the penetration depth λ_0, and the Fermi surface area is estimated.

537.312.62 3066
Superconducting Tunneling on Bulk Niobium—M. D. Sherrill and H. H. Edwards. (*Phys. Rev. Lett.*, vol. 6, pp. 460–461; May 1, 1961.) Measurements of tunneling current for Nb-Pb junctions are described.

537.312.62:539.23 3067
An Upper Limit for the Resistivity of a Superconducting Film—R. F. Broom. (*Nature*, vol. 190, pp. 922–933; June 10, 1961.) A persistent current was induced in a Pb-In film at 2.1°K. Hall-effect probe measurements showed the resistivity to be $<10^{-21}\Omega$.cm.

537.533.8 3068
Reflection of Slow Electrons from Tungsten Single Crystals, Clean and with Adsorbed Monolayers—P. Kisliuk. (*Phys. Rev.*, vol. 122, pp. 405–411; April 15, 1961.) Experimental results are described and considered in relation to theory and previous experimental work. The technique permits continuous recording of the change in work function as gas is adsorbed.

537.533.8:539.23 3069
Certain Regularities in the Secondary Electron Emission from Thin Metal Layers and Semiconductors—I. M. Bronshteĭn and B. S. Fraĭman. (*Dokl. Akad. Nauk SSSR*, vol. 135, pp. 1097–1100; December 11, 1960.) Experimental investigation of the relation between δ and η for thin layers of Ca, Be, Ba or Ti applied respectively to base layers of Be, Ca, Ti, or Ba, where δ and η are the coefficients of slow secondary emission and inelastical reflection.

538.22:538.569.4 3070
Nuclear Quadrupole Resonance in an Antiferromagnet—J. C. Burgiel, V. Jaccarino, and A. L. Schawlow. (*Phys. Rev.*, vol. 122, pp. 429–436; April 15, 1961.)

538.221 3071
The Magnetic Field above the Surface of a Plate-Shaped Domain Structure with Bloch-Wall and Surface Poles—C. Greiner. (*Mber. Dtsch. Akad. Wiss. Berlin*, vol. 2, no. 6, pp. 328–336; 1960.) The field distribution is calculated.

538.221 3072
Magnetic Properties of Fe_3P—M. C. Cadeville and A. J. P. Meyer. (*C.R. Acad. Sci.*, (*Paris*) vol. 251, pp. 1621–1622; October 17, 1960.)

538.221 3073
A Method of Analysis of the Magnetic History of Ferromagnetic Specimens by means of the Preisach Diagram—H. Seidel. (*Z. angew. Phys.* vol. 12, pp. 493–501; November, 1960.) The Preisach diagram [see, *e.g.*, 842 of 1957 (Feldtkeller and Wilde)] can be used to analyze the magnetization processes which a specimen has undergone previously. Discrepancies between theory and measurements relating to Rayleigh's constant are discussed and the use of the principle in the development of storage-type devices is indicated.

538.221 3074
The Influence of Internal Magnetic Couplings on the Shape of Preisach Functions of High-Permeability Materials—H. Girke. (*Z. angew. Phys.*, vol. 12, pp. 502–508; November, 1960.) The unsymmetrical arrangement of Preisach functions already observed are explained [3590 of 1960 (Wilde and Girke)].

538.221 3075
Effects of the Special Heat Treatment on Permalloys—S. Tomita. (*J. Phys. Soc., Japan*, vol. 16, pp. 393–396; March, 1961.) Heat treatment of mumetal and 50 per cent Ni-50 per cent Fe permalloy at temperatures from 200°C to 600°C was found to give a marked increase of maximum permeability. The effect is attributed to inhomogeneous internal structure in these materials.

538.221:538.569.4 3076
Study of the Magnetization in a Single Crystal of Silicon-Iron by Ferromagnetic Resonance—A. Coumes. (*C.R. Acad. Sci., Paris*, vol. 250, no. 21, pp. 3458–3460; May, 1960.) The theory of phases of elementary domains leads to a method of determining the true magnetization direction in a phase from the experimental resonance curve. See 978 of March.

538.221:539.23 3077
Measurements on the Magnetic Anisotropy in Evaporated Iron Films—R. Vrambout and L. De Greve. (*Appl. Sci. Res.*, vol. B9, no. 2, pp. 102–106; 1961.)

538.221:539.23:538.691 3078
The Magnetic Deflection of Electron Beams in Thin Iron Films—H. Boersch, W. Raith, and H. Weber. (*Z. Phys.*, vol. 161, pp. 1–12; December 1, 1960.) Experimental investigations of the effect of film thickness, and the internal magnetic field , on the angle of deflection of an electron beam passing through the film.

538.221:621.318.124:538.569.4 3079
Ferromagnetic-Resonance Line Width in Cobalt-Substituted Ferrites—C. W. Haas and H. B. Callen. (*Phys. Rev.*, vol. 122, pp. 59–68; April 1, 1961.)

538.221:621.318.132 3080
Possibilities of Modifying the Magnetic Properties of Permalloy Alloys—F. Pfeifer. (*Elektrotech. Z., Edn. A*, vol. 81, pp. 945–949; December 1, 1960.) Important characteristics of permalloy-type materials can be modified by high-temperature annealing and by tempering. The results of experimental investigations on various alloys are given and discussed.

538.221:621.318.134 3081
Effect of Boron on Square-Loop Cadmium Manganese Ferrites—B. R. Eichbaum. (*J. Amer. Ceram. Soc.*, vol. 44, pp. 51–54; February 1, 1961.) A ferrite toroid dipped in boron solution before heat treatment possesses much lower coercivity than an undipped core. The switching time, Curie oint and UV_2/UV_2 ratio are also reduced.

538.221:621.318.134 3082
Exchange Anisotropy in Mixed Maganites with the Hausmannite Structure—I. S. Jacobs and J. S. Kouvel. (*Phys. Rev.*, vol. 122, pp. 412–418; April 15, 1961.)

538.221:621.318.134 3083
Magnetic Torque Curves for a Single Crystal of Thulium Orthoferrite ($TmFeO_3$)—C. Kuroda T. Miyadai, A. Naemura, N. Niizeki, and H. Takata. (*Phys. Rev.*, vol. 122, pp. 446–447; April 15, 1961.) Experimental results indicate that a previously proposed model [1956 of 1959 (Sherwood *et al.*)] should be modified.

538.221:621.318.134:548.5 3084
Growth of Single-Crystal Iron Ferrites by the Czochralski Method—F. H. Horn. (*J. Appl. Phys.*, vol. 32, pp. 900–901; May, 1961.)

538.222:538.569.4 3085
Effect of Configuration Mixing and Covalency on the Energy Spectrum of Ruby—S. Sugano and M. Peter. (*Phys. Rev.*, vol. 122, pp. 381–386; April 15, 1961.)

539.23:537.533.7 3086
Hot Electrons in Metal Films: Injection and Collection—J. P. Spratt, R. F. Schwarz, and W. M. Kane. (*Phys. Rev. Lett.*, vol. 6, pp. 341–342; April 1, 1961.) A device is described in which quantum tunneling electrons are emitted into a thin metal film. The structure facilitates the separation of these hot electrons. The characteristics of the device are

given which show that it could be used as an amplifier.

548.5 3087
Radio-Frequency Technique for Pulling Oxide Crystals without employing a Crucible Susceptor—F. R. Monforte, F. W. Swanekamp, and L. G. Van Uitert. (*J. Appl. Phys.* vol. 32 pp. 959–960; May, 1961.) Details are given of a water-cooled RF induction-heating coil which keeps the center part of the charge molten and the outer part solid.

MATHEMATICS

517.512.2 3088
Simplifying the Fourier Integral by Adequate Specification—C. G. Mayo and J. W. Head. (*Brit. J. Appl. Phys.*, vol. 12, pp. 248–250; May, 1961.) Use of a limiting metrical form of definition of the impulse function is shown to be superior to the normal idealized Dirac function in removing ambiguities such as those occurring in an analysis of the voltage developed across a resistance with residual shunt capacity when a current impulse is applied.

517.64:551.510.535 3089
A Solution of the Integral Equation $h'(f)=\int\mu[(f:f_0)dz(f_0)$—Unz. (See 2966.)

517.918 3090
Taylor-Cauchy Transforms for Analysis of Varying-Parameter Systems—Y. H. Ku. (PROC. IRE, vol. 49, pp. 1096–1097; June, 1961.) Six examples are given. See also 2853 of 1960 (Ku *et al.*).

518.4:621.372.5 3091
Graphs with Two Kinds of Elements—S. L. Hakimi. (*J. Franklin Inst.*, vol. 270, pp. 451–467; December, 1960.) Fundamental properties of graphs with red and black elements are studied and the procedures given are modified to find the number of poles and zeros of network functions of an *RLC* network.

519.281.2 3092
A New Technique for Increasing the Flexibility of Recursive Least-Squares Data Smoothing—N. Levine. (*Bell Sys. Tech. J.*, vol. 40, pp. 821–840; May, 1961.) Optimum, or arbitrary, weights can be assigned to the observations, and the restriction of a constant data interval can be removed without affecting the optimum weighting or recursive features.

MEASUREMENTS AND TEST GEAR

529.786:621.3.018.41(083.74) 3093
Results of a Long-Range Clock Synchronization Experiment—F. H. Reder, M. R. Winkler, and C. Bickart. (PROC. IRE, vol. 49, pp. 1028–1032; June, 1961.) Propagation and synchronization experiments are described to examine the possibility of setting up a worldwide network of synthconized atomic clocks. Original synchronization is established by using an atomic clock which is flown from the master to the slave clock and maintaining it by phase tracking a VLF signal controlled by the master clock.

537.311.33:[546.28+546.289 3094
A Rapid and Accurate Method for Measuring the Thickness of Diffused Layers in Silicon and Germanium—B. Jansen. (*Solid-State Electronics*, vol. 2, pp. 14–17; January, 1961.) A cleavage technique is described which permits accurate microscopic measurement of layer thickness.

621.3.018.41(083.7) 3095
Caesium Frequency Standards: Description and Results—P. Kartaschoff, J. Bonanomi, and J. De Prins. (*Helv. Phys. Acta*, vol. 33, no. 9, pp. 968–973; December, 1960. In French.) Brief information concerning the two standards at the Swiss Horological Research Laboratory, Neuchâtel, includes records of frequency comparisons made with other Cs standards from June to September, 1960.

621.317:621.382.3 3096
Transistors in Measurement Techniques: Properties and Application—Schneider. (See 3187.)

621.317.3:621.391.822 3097
Measuring Amplitude and Phase—N. R. Goodman. (*J. Franklin Inst.*, vol. 270, pp. 437–450; December, 1960.) The amplitude and phase of moderately wide-band noise and noise contaminated by extraneous noise can be determined from a pair of finite moving averages on the sample noise record. The method is applicable to radio-star scintillation data.

621.317.3.088 3098
Accuracy in A. C. Measurements—G. H. Rayner and A. Felton. (*J. IEE*, vol. 7, pp. 141–144; March, 1961.) Absolute values of inductance and capacitance can be obtained to an accuracy within 10–20 parts in 10^6, while calibration accuracy is within 30–100 parts in 10^6. Voltage and current at frequencies up to 50 kc can be measured to within 1 part in 10^4, and power and energy to within 5 parts in 10^4.

621.317.313:621.314.224 3099
Techniques for the Calibration of Standard Current Transformers up to 20 kc/s—J. J. Hill. (*Proc. IEE*, pt. B, vol. 108, pp. 333–336; May, 1961. Discussion, pp. 337–338.)

621.317.32:621.375.13.029.4 3100
The Design of an Audio-Frequency Amplifier for High-Precision Voltage Measurement—S. Harkness and F. J. Wilkins (*Proc. IEE*, pt. B, vol. 108, pp. 319–326; May, 1961. Discussion, pp. 337–338.) The amplifier is based on the "ring-of-three" feedback circuit and gives a number of fixed gains between 2 and 1000 in the frequency range 30 cps–30 kc.

621.317.321:537.311.33 3101
Concerning a Rapid Method of Precise Measurements of Thermal E.M.F. of Semiconductors—O. V. Emel'yanenko and F. P. Kesamanly. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1494–1496; July, 1960.) Contact thermocouples with controlled heating are used.

621.317.331:537.311.33 3102
Equipment for Measuring and Recording the Resistance of Semiconductors in Coordinates $\log R=f(1/T)$—L. G. Sapogin and V. M. Ivko. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1482–1488; July, 1960.) The equipment comprises a logarithmic and a hyperbolic amplifier, circuit details of which are given.

621.317.331:537.311.33 3103
An A.C. Bridge for Semiconductor Resistivity Measurements using a Four-Point Probe—M. A. Logan. (*Bell Sys. Tech. J.*, vol. 40, pp. 885–919; May, 1961.)

621.317.335 3104
Improvements in the Precision Measurement of Capacitance—G. H. Rayner and L. H. Ford. (*Proc. IEE*, pt. B, vol. 108, pp. 337–338; May, 1961. Discussion on 2123 of 1960.)

621.317.337:621.372.413 3105
The Effects of Parasitic Modulation on the Accuracy of Measurement of the Q-Factor of a Resonator—F. H. James. (*Proc. IEE*, pt. B, vol. 108, pp. 316–318; May, 1961.) An extension of a previous paper (*ibid.*, pt. B, vol. 106, pp. 489–494; September, 1959) to cover the case where parasitic modulation is found. Theoretical and experimental results in such a case are given.

621.317.343:621.315.212 3106
Impedance Measurement on Coaxial Cables by means of Pulse Echo Tester—R. Codelupi. (*Note Recensioni Notiz.*, vol. 9 pp. 1006–1033; November/December, 1960.)

621.317.361.018.756:621.372.57 3107
The Applicability of Active Filters to the Measurements of the Frequency Spectra of Pulses—Jungmeister. (See 2882.)

621.317.373:621.376.5 3108
Measuring Frequency Stability of Pulsed Signals—R. H. Holman and R. B. Shields. (*Electronics*, vol. 34, pp. 61–65; April 21, 1961.) A method of measuring the phase difference between pulses as a pulse-to-pulse rms deviation, which does not require synchronization or frequency reference from a signal source.

621.317.7:[621.382.3+621.372.5 3109
A Loss and Phase Set for Measuring Transistor Parameters and Two-Port Networks between 5 and 250 Mc/s—D. Leed and O. Kummer. (*Bell Sys. tech. J.*, vol. 40, pp. 841–884; May, 1961.) The maximum inaccuracy is 0.1 db in loss, and 0.5° in phase. The special problems of broad-band measurement on transistors are reviewed, and details are given of the experimental equipment and the results obtained.

621.317.7:621.397.6 3110
Some New Video Measurement Techniques and Apparatus—Weaver. (See 3166).

621.317.74 3111
A Transfer-Function Analyser for Linear and Nonlinear Components—A. K. Choudhury, M. S. Basu and A. K. Mahalanabis. (*Electronic Eng.*, vol. 33, pp. 382–385; June, 1961.) A method of measuring frequency response, gain and phase is described requiring only normal laboratory components.

621.317.75 3112
A Voltage Waveform Analyzer—R. Dehors and G. Seguier. (*C.R. Acad. Sci. Paris*, vol. 250, pp. 3464–3466; May 23, 1960.) A bridge method of analysis is described in which voltage measurements can be made at any point in the cycle.

621.317.75 3113
Phase Sensitive Wave Analysis—O. J. Haga and D. Midgley. (*Electronic Tech.*, vol. 38, pp. 257–261; July, 1961.) "A method of phase analysis is described in which the input waveform is applied to one input of a multiplier. The other input takes in turn reference sine and cosine waves generated locally at the harmonic frequencies. To maintain phase-coherence, operation is interrupted in alternate fundamental periods, allowing initial conditions to be reset in the reference-wave generator. Results are given for an experimental circuit with a wattmeter performing the function of multiplication."

621.317.79.029.64/.65 3114
A Microwave Interferometer with Coherent Background—A. Boivin and R. Tremblay. (*Canad. J. Phys.*, vol. 39, pp. 393–408; March, 1961. In French.) The operation of the instrument is based on the principle of coherent background. The reference signal is distributed over the whole of the image space, using suitably polarized sources and a turnstile antenna for detection. The diffraction pattern is scanned twice to determine both the real and imaginary

parts, in Cartesian form, of the distribution of complex amplitude.

621.317.794:523.164 3115
Digital Radiometer—S. Weinreb. (PROC. IRE, vol. 49, p. 1099; June, 1961.) The power spectrum of nosmic noise is measured, over a 5-Mc band, via its autocorrelation function. To eliminate the effects of gain fluctuation, the signal is amplified and limited; the autocorrelation of the resulting square waves is found by a digital technique and, assuming a Gaussian input signal, is transformed to the power spectrum. It is hoped that one can detect the galactic deuterium line.

621.317.794:621.396.62.029.65/.66 3116
Submillimetre-Wave Radiometry—Long and Rivers. (See 3143.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

562.2:621.396.9 3117
An Electronic Distance Measurement System for Accurate Survey—(*Proc. IRE Australia*, vol. 22 pp. 22–25; January 1961.) An explanation of the principles and operation of the "Tellurometer" relating to its application in Australia. See 1980 of 1959 (Wadley) and back references.

535.376.07:537.227 3118
Ferroelectric Scanning of Electroluminescent Displays—M. Cooperman. (*RCA Rev.*, vol. 22 pp. 195–205; March 1961.) A new scanning method using a $BaTiO_3$ crystal is described.

536.587:621.382.3 3119
The Transistor as a Temperature-Sensing Device in Temperature Control Systems—J. E. Pallett. (*Electronic Eng.* vol. 33 pp. 360–363; June, 1961.) A detailed examination is made of the use of a transistor in a common-base circuit for the detection of small temperature changes.

537.533:621.9 3120
The Electron Beam as a Tool—S. Panzer and K. H. Steigerwald. (*Elektrotech. Z., Edn A*, vol. 81, pp. 925–932; December 19, 1960.) The techniques reviewed include welding, soldering, cutting, milling, drilling, alloying and vapor deposition.

621.362:621.387 3121
Calculation of the Maximum Efficiency of the Thermionic Converter—J. H. Ingold. (*J. Appl. Phys.*, vol. 32, pp. 769–772; May, 1961.)

621.362:621.387 3122
Oscillations and Saturation-Current Measurements in Thermionic Conversion Cells—R. J. Zollweg and M. Gottlieb. (*J. Appl. Phys.*, vol. 32, pp. 890–894; May, 1961.) RF oscillations in Cs-filled thermionic diodes are interpreted on the basis of a model which assumes the ions oscillate in an excess negative charge potential well outside the cathode.

621.387.462:537.311.33 3123
On the Energy Expended per Electron-Hole Pair Produced in p-n-Junction Detectors—E. Baldinger and W. Czaja. (*Nuclear Instr. Meth.*, vol. 10, pp. 237–239; March, 1961.) Previous work relating to radiation counters [see, *e.g.*, *Proc. Roy. Soc.* (*London*), *A*, vol. 224, no. 1158 pp. 362–373; July, 1954 (Erskine)] indicates that for gases, the energy loss $\epsilon = k_1 + k_2 I$, where k_1 and k_2 are constants and I is the ionization energy of the gas. A similar correlation holds for semiconductors. See also 2753 of August (Baldinger *et al.*).

PROPAGATION OF WAVES

621.391.812.62 3124
Experimental Results with Transhorizon Tropospheric Propagation—F. du Castel. (*Ann. Télécommun.*, vol. 15, pp. 255–259; November/December, 1960.) Results of experimental studies made in the period 1957–1960 are discussed.

621.391.812.62 3125
Theories of Tropospheric Propagation Beyond the Horizon—J. Voge. (*Ann. Télécommun.*, vol. 15, pp. 260–265; November/December, 1960.) A review, with 38 references.

621.391.812.62 3126
Several Aspects of Radio-Meteorology and Radio-Climatology—P. Misme. (*Ann. Télécommun.*, vol. 15, pp. 266–273; November /December, 1960.) Three aspects are considered: 1) investigation of a model atmosphere capable of sustaining transhorizon propagation, 2) refraction, 3) choice of parameters in radioclimatology.

621.391.812.62:551.510.62 3127
Radio-climatology Tests in the Congo Basin—P. Misme. (*Ann. Télécommun.*, vol. 16, pp. 28–40; January/February, 1961.) An investigation of the refractive index of the air, its gradient, the stability of the lower atmosphere and the water-vapor pressure at different heights.

621.391.812.62.029.63 3128
Some Factors Influencing 3-cm Radio-Wave Propagation Oversea Within and Beyond the Radio Horizon—F. A. Kitchen, W. R. R. Joy, and E. G. Richards. (*Proc. IEE*, pt. B, vol. 108, pp. 257–263; May, 1961.) Propagation of 10-Gc radio waves over sea did not agree with calculations based on an atmosphere of uniform refractive index gradient. Signal losses of 5 – 3C db often occurred within the horizon and were usually accompanied by an interference pattern.

621.391.812.62.029.64 3129
Experimental Investigation of the Instantaneous Frequency Structure of Propagation in a Heterogeneous Medium—J. Biggi, F. du Castel, J. C. Simon, A. Spizzichino, and J. Voge. (*Ann. Télécommun.*, vol. 15, pp. 274–276; November/December, 1960.) Experimental results obtained over a 300-km radio link are given. The transmitter frequency is swept over the range 3.3–3.5 Gc in 1/20 sec and the received signal analyzed as a function of either amplitude or time.

621.391.812.63 3130
The Fading of Radio Waves Weekly Scattered at Vertical Incidence from Heights near 90 km—O. Awe. (*J. Atmos. Terr. Phys.*, vol. 21, pp. 142–156; June, 1961.) At 2–3 Mc the fading rate was about 20 maxima/minutes. The echo shape is often such as would be expected from a rough screen. Appreciable energy is returned from directions well away from the vertical. The fading is due to random movements of the scattering elements whose rms velocity is about 13 m/sec.

621.391.812.63 3131
The Fading of Radio Waves Reflected from the Ionosphere at Oblique Incidence—O. Awe. (*J. Atmos. Terr. Phys.* vol. 21, pp. 120–141; June, 1961.) Pusle transmissions near 2 Mc were used. Echoes from 90 km and 120 km had fading periods near 1 minute and 6 sec respectively. Diffraction studies suggest that the irregularities near 90 km are horizontal disks while those at 120 km are spherical. Those near 300 km are elongated along the earth's field with an axial ratio of 3:1.

621.391.812.63 3132
Results of the Approximate Calculation of Space-Wave Field Strength at the Limits of the Transmission-Frequency Range—B. Beckmann. (*Nachrtech. Z.*, vol. 13 pp. 470–472; October, 1960.) Examples of the application of the formula given in 306 of 1960 are discussed.

621.391.812.63.029.45 3133
Effect of Underground Induced Polarization on E.L.F. Propagation—H. Raemer. (*J. Geophys. Res.*, vol. 66, pp. 1596–1597; May, 1961.) The largest possible corrections to the mode parameters are found to be negligible.

621.391.812.63.029.45 3134
A Comparison between Theoretical Data on Phase Velocity of VLF Radio Waves—J. R. Wait. (PROC. IRE, vol. 49, pp. 1089–1090; June, 1961.) Fairly good agreement is indicated between the measured phase velocity and mode theory. The ionosphere is assumed to be sharply bounded, with its lower edge at 70 km during the day and 90 km at night.

621.391.812.63.029.45 3135
The Numerical Solution of Differential Equations governing the Reflexion of Long Radio Waves from the Ionosphere: Part 4—D. W. Barron. (*Proc. Roy. Soc.* (*London*) *A*, vol. 260, no. 1302, pp. 393–408; March, 1961.) Reflexion coefficients calculated for a series of models of the daytime ionosphere in summer by methods described previously, are compared with observed values. The model whose theoretical properties most nearly agree with experimental observations is found and suggestions are made as to how it should be changed. Sunrise and sunset transitions are also discussed and it is shown that the sharp change in reflection coefficients can be explained in terms of a low-level D layer which is only present in the daytime. Part 3: 3455 of 1959 (Barron and Budden).

621.391.812.63.029.62 3136
Studies of Ionospheric Forward Scattering using Measurements of Energy Distribution in Azimuth—W. C. Bain. (*Proc. IEE*, pt. B, vol. 108, pp. 241–252; May, 1961.) A marked diurnal variation of the mean bearing of the 37-Mc forward-scatter signal was found. This and other results suggested that the scattering process on the 1740-km link used was due mainly to meteor reflections at night but that during the day turbulent scattering was also of importance.

621.391.812.63.029.62 3137
Directional Observations on Delayed Signals on an Ionospheric Forward-Scatter Circuit—W. C. Bain. (*Proc. IEE*, pt. B, vol. 108, pp. 253–256; May, 1961.) A study was made of the mean bearing of the delayed signals arriving at Slough from Gibraltar after reception of the forward-scatter signal. Results for the first two of these to arrive at Slough are given, and also details of round-the-world echoes. See also 3136 above.

621.391.812.63.029.63 3138
The Effect of Faraday Rotation on Incoherent Back-Scatter Observations—G. H. Millman, A. J. Moceyunas, A. E. Sanders, and R. F. Wyrick. (*J. Geophys. Res.*, vol. 66, pp. 1564–1568; May, 1961.) Preliminary observations of incoherent scattering conducted at Trinidad (10.7°N, 61.6°W) with a high-power pulse radar operating at 400 Mc are described.

RECEPTION

621.376.23 3139
Certain Problems concerning the the Detection of Weak Signals—B. Picinbono. (*Ann. Télécommun.*, vol. 16, pp. 2–27; January/Feb-

ruary, 1961.) A thesis; for shorter papers on two different aspects of this work, see 1630 of May and 2763 of August.

621.391.82:621.396.73 3140
Probabilities of Interference with Mobile Field Radio Derived from a Field-Strength Survey at 59 Mc/s—D. R. W. Thomas. (*Proc. IEE*, pt. B, vol. 108, pp. 264–272; May, 1961.) The variation of field strength at 59 Mc with distance was found to agree well with calculations, its distribution at a given distance was found to be log-normal. Probabilities of interference are derived and presented graphically.

621.391.823 3141
Radio Interference from Ignition Systems—A. H. Ball and W. Nethercot. *Proc. IEE*, pt. B, vol. 108, pp. 273–278; May, 1961.) Comparison of American, German and British measuring equipment, techniques and limits. Good agreement was found between the various measuring sets.

621.391.827 3142
A Mechanism for Direct Adjacent-Channel Interference—C. L. Ruthroff. (Proc. IRE, vol. 49, pp. 1091–1092; June, 1961.) The effects of limiters can cause the modulation in a FM channel to appear as undistorted modulation in an adjacent channel.

621.396.62.029.65/.66:621.317.794 3143
Submillimetre-Wave Radiometry—M. W. Long and W. K. Rivers, Jr. (Proc. IRE, vol. 49, pp. 1024–1027; June, 1961.) Greater sensitivity in reception at sub-mm λ will result from the use of a direct detection system using a chopper, square-law detector and narrow-band amplifier than will be obtainable with a superheterodyne system.

STATIONS AND COMMUNICATION SYSTEMS

621.376.56:621.391.82 3144
Errors in Detection of R.F. Pulses Embedded in Time Crosstalk, Frequency Crosstalk, and Noise—E. A. Marcatili. (*Bell Sys. Tech. J.*, vol. 40, pp. 921–950; May, 1961.) Calculations are given of the probability of error in the detection of RF pulses in a combination of Gaussian noise, time crosstalk from two neighboring pulses, and frequency crosstalk from an adjacent channel. Criteria for the optimum design of a PCM system are discussed.

621.376.56:621.391.82 3145
Time and Frequency Crosstalk in Pulse-Modulated Systems—E. A. Marcatili. (*Bell Sys. Tech. J.*, vol. 40, pp. 951–970; May, 1961.) The time and frequency crosstalk between Gaussian RF pulses sent via adjacent frequency channels over the same transmission medium is calculated. See also 3144 above.

621.396.41 3146
A Study by means of an Analogue Computer of the Spectrum of Compatible Single-Sideband Modulation—H. Mertens. (*E.B.U. Rev.*, no. 64A, pp. 249–258; December, 1960.) Analysis of the signal spectrum produced by the Kahn system of SSB modulation (971 of 1959).

621.396.43:551.507.362.2 3147
Application of Artificial Satellites to Communications on Ultra Short Waves—M. Dolukhanov. (*Radio, Moskow*, no. 5, pp. 21–22; May, 1961.) A note on satellite period and transmitter power for heights of 1000, 2000, 3000, 4000, 5000 and 35,818 km.

621.396.43:551.507.362.2 3148
A Transatlantic Communication Experiment via Echo I Satellite—W. C. Jakes, Jr. (*Nature*, vol. 190, p. 709; May 20, 1961.) Further details are given relating to the experiment described in 1306 of April (Carru *et al.*).

621.396.43:551.507.362.2:621.396.677.3 3149
Communications Satellites Using Arrays—R. C. Hansen. (Proc. IRE, vol. 49, pp. 1066–1074; June, 1961.) Automatic angle return arrays are investigated for both passive and active systems. These Van Atta arrays return a signal in the direction of incidence and are effective over at least $\pm 45°$. Thus only partial stabilization or single axis (spin) stabilization need be used, greatly simplifying the station-keeping orbit control problem. An active Van Atta scheme has inherently high reliability since many of the distributed amplifiers can fail without serious performance degradation. The distributed structure also allows use of low-power solid-state amplifiers.

621.396.65.029.63 3150
FM120/2200: Radio-Link Installation for 120 Telephony Channels in the 2-GC/s Band—(*Telefunken Ztg.*, vol. 33, pp. 295–323; December, 1960.)

Part 1: Radio-Link System—E. Willwacher (pp. 295–302, English summary, pp. 334–335).

Part 2: Modulation Equipment—R. Heer and H. Oberbeck (pp. 302–305, English summary, p. 335).

Part 3: Radio Equipment—W. Schlotterbeck (pp. 306–309, English summary, p 335).

Part 4: Design of Aerial Duplexers—W. Schlotterbeck and E. Willwacher (pp. 309–313, English summary, p. 336).

Part 5: Stand-By Circuit—J. Goeldner and O. Tegel (pp. 313–318, English summary, p. 336).

Part 6: Equipment for Diversity Reception—H. Weber (pp. 318–323, English summary, p. 336).

621.396.712 3151
B.B.C. Sound Broadcasting 1939–60—E. L. E. Pawley. (*Proc. IEE*, pt. B, vol. 108, pp. 279–302; May 1961.) A review of progress.

621.396.946 3152
Frequency Allocations for Space Communications—(Proc. IRE, vol. 49, pp. 1009–1015; June, 1961.) The technical aspects of space communications are examined with special attention to the use of satellites for commercial trunking purposes. The influence of system design on frequency allocation and utilization, and associated problems are discussed.

SUBSIDIARY APPARATUS

651–52 3153
Theory of Nonlinear Control—Y. H. Ku. (*J. Franklin Inst.*, vol. 271, pp. 108–144; February, 1961.) A review of work in various fields of nonlinear control (157 references).

621.311.62:621.383.292 3154
Stabilized E.H.T. Supply Unit for Photomultipliers—H. Constans and D. Vallat. (*Electronic Appl.*, vol. 20, pp. 153–158; February, 1961.) Provided the load is substantially constant, the circuit will produce an output voltage of 1000–2000v with an error not exceeding 0.5 per cent during mains fluctuations of ± 10 per cent.

621.311.69:621.383.5 3155
Spectral Response of Solar-Cell Structures—L. M. Terman. (*Solid-State Electronics*, vol. 2, pp. 1–7; January, 1961.) Measurements on Si cells with the *p-n* junction 0.6 to 5.0 μ below the surface show that the short wave response ($\lambda > 0.75$ μ) is relatively increased by making the junction close to the surface, and the long-wave response by making it deeper. Effects of rough and smooth surfaces were also studied.

621.314.63 3156
On the Electrical Parameters of some Types of Selenium Rectifier—I. Kh. Geller and P. V. Sharavskii. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1441–1449; July, 1960.)

621.316.72:621.318.3 3157
Variable, Precision-Regulated, Low-Voltage High-Current Supply for Large Electromagnets—R. C. Mobley. (*Rev. Sci. Instr.*, vol. 32, pp. 432–433; April, 1961.) Circuit details are given of a power supply with combined electronic- and magnetic-amplifier control delivering currents of 5–200 A at 40v with regulation and ripple values <0.01 per cent.

TELEVISION AND PHOTOTELEGRAPHY

621.397.12:621.397.232.6 3158
A New Type of Vestigial-Sideband Facsimile System K. Kubota and K. Kobayashi. (*Rev. Elect. Commun. Lab., Japan*, vol. 9, p. 85–90; January/February, 1961.) A simple system performing homodyne detection and eliminating quadrature distortion, which does not require AFC.

621.397.13:621.395.625.3 3159
A Magnetic Wheel Store for Recording Television Signals—J. H. Wessels. (*Philips Tech. Rev.*, vol. 22, pp. 1–10; November 4, 1960.) The device described will store one or two frames of a television signal indefinitely. The recording is made on the periphery of a magnetically coated rotating wheel using a frequency-modulation technique. A useful application in radiology is mentioned.

621.397.132 3160
Comparative Investigations of Modified N.T.S.C. Colour Television Systems using Pre- or Post-correction of Gradation—K. Bernath. (*Rundfunktech. Mitt.*, vol. 4, pp. 232–237; December, 1960.) The advantages of pre-correction are established on the basis of experimental results.

621.397.132 3161
Colour Television Transmission with Simultaneous Frequency and Amplitude Modulation of the Colour Carrier (F. A. M. Method)—N. Mayer. (*Rundfunktech. Mitt.*, vol. 4, pp. 238–252; December, 1960.) The system described combines some of the advantages of the "Henri de France" (*SECAM*) system of color television (see, *e.g.*, 3660 of 1960) with those of the NTSC method. Pictures obtained on monochrome and color television receivers are reproduced.

621.397.132 3162
Trends towards a Common European Standard for Colour Television—F. Cappuccini. (*Note Recensioni Notiz.*, vol. 9, pp. 993–1005; November/December, 1960.) The problem of establishing a standard for compatible color television is related to the unification of European monochrome standards. Particular consideration is given to a 625-line standard with 8-Mc channel width.

621.397.331:621.391 3163
Television Band Compression by Contour Interpolation—D. Gabor and P. C. J. Hill. (*Proc. IEE*, pt. B, vol. 108, pp. 303–315; May, 1961.) This method of compression uses redundancies between fields and between frames. The waveband gain is 4:1 without deterioration in picture quality. The results of tests in a photomechanical model are given.

621.397.331.24 3164
Measurements of Sharpness on Picture Tubes—W. Schröder. (*Telefunken-Röhre*, no. 37, pp. 17–70; October, 1960.) A method is described which permits the measurement of fo-

cusing quality at any point on the screen of a television tube. The various factors influencing spot size are discussed with reference to the results of measurements.

621.397.334.2:778.6 3165
Electronic Reversal of Photographic Colour Negatives—K. Welland. (*Arch. elekt. Übertragung*, vol. 14, pp. 441–450; October, 1960.) An experimental flying-spot scanner with three input channels and a display unit is described. Correction and calibration methods are discussed. Color reproductions of photographically and electronically reversed test pictures are given for comparison.

621.397.6:621.317.7 3166
Some New Video Measurement Techniques and Apparatus—L. E. Weaver. (*E.B.U. Rev.*, no. 64A, pp. 242 248; December, 1960.) Description of techniques and apparatus used by the BBC, including a sine-squared pulse and bar generator, a test-waveform generator for nonlinear distortion measurements, and an outline of the spectrum method of measuring random noise (2345 of 1959). For German version see *Rundfunktech. Mitt.*, vol. 5, pp. 8–14; February, 1961.

621.397.61 3167
Methods and Limitations of Television Camera Techniques—R. Theile. (*Elektrotech. Z., Edn A*, vol. 81, pp. 895–903; December 5, 1960.) The principles underlying electron-optical image conversion are outlined and flying-spot scanning and storage-type camera-tube techniques are discussed in greater detail. Picture quality and sensitivity limits are considered in relation to results obtained by photographic means.

621.397.61:621.396.677.3 3168
Dipole Arrays for Transmitter Aerials for Television Bands IV/V—Laub. (See 2854.)

621.397.62 3169
Synchronizing and A.G.C. Circuits in TV Receivers—E. M. Cherry. (*Proc. IRE, (Australia)*, vol. 22, pp. 61–76; February, 1961.) Practical details are given for the design of a high-quality television receiver from fundamental considerations. The design is simple and noncritical.

621.397.62:616–001.26 3170
Aspects of the Emission of X Rays from Television Receivers—M. V. Callendar and D. F. White. (*J. Brit. IRE*, vol. 21, pp. 389–400; May, 1961.) Measurements made with a proportional counter of the dose rates from typical picture tubes show the radiation level is considerably less than the recommended level of 0.5 milliröntgens per hour.

621.397.62.004.6 3171
Component and Valve Reliability in Domestic Radio and Television Receivers—D. W Heightman. (*J. Brit. IRE*, vol. 21, pp. 401–407; May, 1961. Discussion.) Statistical information on the reliability of television receivers is reviewed. Some reasons for failure are examined and suggestions for improvements made.

621.397.63:621.395.625.3 3172
A Translator-Modulator-Demodulator as used in V.H.F./F.M. Techniques, for Television Tape-Recording Equipment—H. Fix and W. Habermann. (*Rundfunktech. Mitt.*, vol. 4, pp. 222–231; December, 1960.) The modification of the Ampex method of video tape recording to meet the requirements of the CCIR 625-line standard is discussed. The use of double frequency-changing overcomes most of the quality defects.

621.397.7 3173
Television Centre of Greater Moscow—(*Radio, Moskow*, no. 1, pp. 6–7; January, 1961.) Brief description of a broadcasting tower to be built in Moscow which will have a base diameter of 65 m and a height of 520 m. The station will transmit simultaneously six sound and five television programmes, one being in color, over a radius of 120–130 km.

TRANSMISSION

621.311.61:551.507.362.2 3174
Beacon Transmitters and Power Supply for Echo I—J. G. McCubbin and H. B. Goldberg. (*RCA Rev.*, vol. 22, pp. 147–161; March, 1961.) The transmitter has an output of 5–10 mw at 107.94 Mc and is mounted with Si solar cells, batteries and antenna on a printed-circuit board 10 inches in diameter.

621.396.61:621.396.662.6 3175
The Design of a Highly-Stable Short-Wave Exciter Stage with Decade Switching in the Range 1.5–30 Mc/s—H. Valdorf and R. Klinger. (*Frequenz*, vol. 14, pp. 335–343; October, 1960.) A crystal-controlled oscillator provides the standard frequency from which the output frequencies are derived by multiplication, mixing and division. Decade switching in 1-kc steps and facilities for interpolation are available in the unit described.

TUBES AND THERMIONICS

621.382.2 3176
Properties and Applications of Semiconductor Diodes—H. F. Grave. (*Elektrotech. Z., Edn A*, vol. 81, pp. 761–767; October 24, 1960.) Circuits for class-B and class-C demodulation are given, and diode applications in the field of HF and h.v. measurement, and as protective and control devices are briefly reviewed.

621.382.2 3177
Investigation of the Characteristics of Germanium Diodes—I. M. Senderikhin and P. V. Sharavskiĭ. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1497–1505; July, 1960.) Static and pulse characteristics of thermistors and Ge point-contact and junction diodes are discussed in relation to thermal effects.

621.382.2 3178
Contribution to the Theory of Thermal Breakdown of Germanium Diodes I. M. Senderikhin. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1506–1517; July, 1960.) The reverse characteristics of Ge point-contact and junction diodes are analyzed and general theory of thermal breakdown is proposed. In point-contact diodes breakdown occurs earlier due to the field effect.

621.382.2 3179
Field-Effect Pinch-Off of Surface Leakage in High-Voltage Diodes—T. Kan. (*Solid-State Electronics*, vol. 2, pp. 68–69; January, 1961.) It is shown how reduced surface leakage and increased reverse-breakdown voltage can be achieved by proper geometry, so as to utilize the field-effect pinch-off.

621.382.22 3180
The Electrical Properties of Extensive Current Paths in Space-Charge Regions Close to the Surface of Semiconductors—E. Groschwitz, E. Hofmeister, and R. Ebhardt. (*Z. angew. Phys.*, vol. 12, pp. 544–557; December, 1960.) Theoretical treatment of surface currents in inversion layers, particularly relating to conditions in point-contact semiconductor diodes. See also 2799 of August.

621.382.23 3181
Construction and Characteristics of Silver-Bonded Diodes—S. Kita. (*Rev. Elect. Commun. Lab., Japan*, vol. 9, pp. 26–30; January/February, 1961.)

621.382.23 3182
High-Frequency Power in Tunnel Diodes—G. Dermit. (Proc. IRE, vol. 49, pp. 1033–1042; June, 1961.) An expression is developed for HF tunnel-diode power with a simple dissipative load and for signals confined within the nearly linear range of the diode negative resistance. Results are plotted for Ge and GaAs.

621.382.23 3183
The Equivalent Noise Current of Esaki Diodes—R. A. Pucel. (Proc. IRE, vol. 49, pp. 1080–1081; June, 1961.)

621.382.23 3184
Gallium Antimonide Esaki Diodes for High-Frequency Applications—C. A. Burrus. (Proc. IRE, vol. 49, p. 1101; June, 1961.) Diodes have been made, using both *p*- and *n*-type material, by methods similar to those applied for GaAs (*J. Appl. Phys.*, vol. 32, pp. 1031–1036; June, 1961.). They have been tested as oscillators in the 50-Gc range.

621.382.23:621.372.44 3185
Some Effects of Material Parameters on the Design of Surface-Space-Charge Varactors—D. R. Frankl. (*Solid-State Electronics*, vol. 2, pp. 71–76; January 1961.) Theoretical discussion of the effects of material parameters (semiconductor and dielectric materials and layer thicknesses) on the capacitance variation and cut-off frequency of varactor diodes.

621.382.23:621.375.9 3186
A Millimetre-Wave Esaki-Diode Amplifier—C. A. Burrus and R. Trambarulo. (Proc. IRE, vol. 49, pp. 1075–1076; June, 1961.) Amplification with Esaki diodes has been achieved in the frequency range 55–85 Gc. Gain was 20 db with 40 Mc bandwidth. Noise figure was 16–18 db, mainly due to circuit losses.

621.382.3:621.317 3187
Transistors in Measurement Techniques: Properties and Application—H. Schneider. (*Elektrotech. Z., Edn A*, vol. 81, pp. 767–773; October, 1960.) Analysis of the properties and characteristics of transistors to assess their suitability for various measurement applications; the advantages and disadvantages in comparison with relays and thermionic tubes are indicated.

621.382.3:621.318.57 3188
Storage Time of a Transistor with a Decaying Turn-Off Current—D. M. Taub. (*Proc. IEE*, pt. B, vol. 108, pp. 344–347; May, 1961.) A theoretical calculation of the storage time.

621.382.3.002.2 3189
Methods of Producing Stable Transistors—J. J. A. Ploos van Amstel. (*Philips Tech. Rev.*, vol. 22, pp. 204–214; March 14, 1961.) Very stable Ge transistors are produced by "surface forming" at 140° C, with the transistor encapsulated in a mixture of silicone grease, boracic acid and arsenic.

621.383 3190
Photoelectric Cells and Photomultipliers—J. Sharpe. (*Electronic Tech.*, vol. 38, pp. 196–201 and 248–256; June and July, 1961.) A review of the properties and applications of photoemissive cells and photomultiplier tubes.

621.383.5 3191
The Influence of the Surface Recombination Velocity and the Absorption Coefficient on the Transient Characteristics of Photodiodes—A. A. Grinberg and N. B. Strokan. (*Fiz.*

Tverdogo Tela, vol. 2, pp. 1536–1541; July, 1960.)

621.385.032.213.23 3192
Emission of Negative Ions of Oxygen from Dispenser Cathodes: Parts 1 & 2—N. A. Surplice. (*Brit. J. Appl. Phys.*, vol. 12, pp. 214–219 and 220–221; May, 1961.) Report of experimental investigations of negative-ion emission from cathodes of BaO in sintered Ni and of sintered W impregnated with Ba aluminate.

621.385.032.213.23 3193
The Emission Equation of the Oxide Cathode—J. Ruf. (*Telefunken—Röhre*, no. 37, pp. 79–132; October, 1960.) Thermionic-emission theory is reviewed and emission equations relating to various models of oxide cathode are given (46 references).

621.385.032.213.23 3194
The Evaluation of Oxide-Cathode Quality by Shot-Noise Tests—L. G. Sebestyen. (*J. Brit. IRE*, vol. 21, pp. 463–467; May, 1961.) The shot-noise test appears to be more sensitive than other methods and does not affect the cathode. Simple test equipment is described. See also 4018 of 1958 (Dahlke and Dlouhy).

621.385.2 3195
The Effect of Elastically Reflected Electrons on the Characteristics of a Thermionic Diode—K. H. V. Booth and I. A. Harris. (*Proc. Roy. Soc. (London) A*, vol. 261, no. 1304, pp. 134–152; April, 1961.) The elastic reflection of slow electrons (<2 ev) in a diode with an oxide-coated cathode has a major effect on the I/v characteristic and the electric field distribution.

621.385.3:621.396.61 3196
Calculation of Optimum Operational Data of Transmitter Valves—G. Schuster. (*Hochfrequenztech. u. Elektroakust.*, vol. 69, pp. 124–132; August, 1960.) Calculations and characteristic curves refer to the use of a power triode in induction heating equipment, but are of general application in determining optimum operating conditions.

621.385.3.029.6 3197
Applications of Microwave Triodes—J. P. M. Gieles. (*Philips Tech. Rev.*, vol. 22, pp. 16–28; November 4, 1960.) A comprehensive survey of current practice in the commercial field.

621.385.6:537.533.08 3198
Electron-Beam Analyzer using a Piezoelectric Scanner—T. Fujii. (*Rev. Sci. Instr.*, vol. 32, pp. 434–444; April, 1961.) This instrument measures transverse current density distributions of high-density electron beams using a moving pinhole target driven by a pair of $BaTiO_3$ elements. Sampling can be carried out very close to the anode aperture of an electron gun

621.385.6:621.375.9:621.372.44 3199
A Low-Noise Microwave Quadrupole Amplifier—A. Ashkin. (Proc. IRE, vol. 49 pp. 1016–1020; June, 1961.) The design of a quadrupole amplifier is discussed and experimental results using a frequency of 4137 Mc are described. At a gain of 19 db the measured double-channel noise figure is 0.79 db.

621.385.62/.63:621.391.832.4 3200
New Results on Frequency Multiplication and Nonlinear Phase Distortion in Klystrons and Travelling-Wave Tubes—F. Paschke. (*RCA Rev.*, vol. 22, pp. 162–184; March, 1961.) The nonlinear space-charge-wave equation is solved by third-order successive approximation. Phase delay and frequency multiplication effects agree well with experimental observations.

621.385.623.5 3201
Effect of Magnetic Fields upon Reflex-Klystron Characteristics—M. E. Brodwin and J. W. Davis. (*Rev. Sci. Instr.*, vol. 32, pp. 223–224; February, 1961.) Experimental results show that magnetic control of frequency is as effective as repeller control, and that the controlling elements are independent of the klystron power supply. This system should be advantageous for the stabilization of high-voltage mm-wave klystrons.

621.385.63 3202
Power Carried by the Cyclotron Waves and the Synchronous Waves on a Filamentary Electron Beam—S. Saito. Proc. IRE, vol. 49, pt. 1, pp. 969–970; May, 1961.)

621.385.63:621.373.423 3203
An Electrically Tunable Microwave Oscillator with High Efficiency and a Constant Output Level Independent of Frequency—W. Eichin and H. Heynisch. (*Nachrtech. Z.*, vol. 13, pp. 457–461; October, 1960.) The oscillator described consists of a combination of a backward-wave oscillator with a traveling-wave amplifier system in the same envelope and coupled by a common electron beam. The efficiency obtainable is that of a traveling-wave tube. Experimental results are given which were obtained with modified traveling-wave tubes containing a section of interdigital line or an eccentrically arranged helix, operating in the 4–6-Gc region.

621.385.63:621.375.9:621.372.44 3204
Excitation and Amplification of Cyclotron Waves and Thermal Orbits in the Presence of the Space Charge—R. Adler, A. Ashkin, and E. I. Gordon. (*J. Appl. Phys.*, vol. 32, pp. 672–675; April, 1961.) The effect of space-charge on the motion of an electron beam and its individual electrons in the transverse fields of cyclotron-wave amplifiers is considered. The effect is such that low-field quadrupolar amplifiers do not saturate at low gains.

621.385.63:621.375.9:621.372.44 3205
Travelling-Wave Analysis of a Class of Parametric Amplifiers Based upon the Hill Equation—R. W. Fredricks. (*J. Appl. Phys.*, vol. 32, pp. 901–904, May, 1961.) Formulas are deduced for the complex wave number of beams, and conditions for growing-wave solutions are presented. The case of a sinusoidal pump field variation is treated, the optimum condition obtains when the pump frequency is twice the beam natural frequency.

621.385.632:621.391.822 3206
Noise Smoothing by Reactive Damping in Finite Multi-velocity Electron Beams—J. Berghammer. (*RCA Rev.*, vol. 22, pp. 185–194; March, 1961.) In an electron beam with rectangular velocity distribution a pair of reactively damped waves exist in addition to the Hahn-Ramo waves. Noise smoothing can result in long low-velocity beams of high density.

621.385.632.12 3207
Propagation of E.M. Waves along a Helix—R. Dingeldey and A. Cunliffe. (*Electronic Tech.*, vol. 38, pp. 262–266; July, 1961.) The introduction of an inner conductor along the axis of a helix to provide electrostatic focusing in a traveling-wave tube is considered. The inner conductor allows both slow and fast waves to be propagated corresponding respectively to the waves propagated along the helix and the conductor. Each is modified by the presence of the other. Experimental data are discussed.

621.385.69:621.372.2 3208
Distributed Amplifier Tube—T. Kojima. (*Rev. Elect. Commun. Lab., Japan*, vol. 8, pp. 343–368; July/August, 1960.) The anode and grid are helical lines with a common velocity of propagation for EM waves. Theory of amplification and measured characteristics are given for two experimental tubes, one of which has an operating range 0–150 Mc. See 1081 of 1960.

621.385.832 3209
Dynamic Deflection Sensitivity of Inclined Deflection Plates—E. Gundert. (*Telefunken—Röhre*, no. 37, pp. 5–16; October, 1960.) The deflection angle of an electron beam at the end of inclined deflection plates is calculated for frequencies at which electron transit time is no longer negligible with respect to the period of the deflection voltage. The plate length for the largest deflection angle is also calculated.

621.385.832:681.142 3210
Design of a High-Resolution Electrostatic Cathode-Ray Tube for the Flying-Spot Store—H. G. Cooper. (*Bell Sys. Tech. J.*, vol. 40, pp. 723–759; May, 1961.) The design is discussed in relation to optimum gun performance and minimum deflection focusing. A substantial reduction is effected in beam aberrations due to fringing of the deflection fields.

621.387:621.362 3211
Calculation of the Maximum Efficiency of the Thermionic Converter J. H. Ingold. (*J. Appl. Phys.*, vol. 32, pp. 769–772; May, 1961.)

621.362:621.387 3212
Oscillations and Saturation-Current Measurements in Thermionic Conversion Cells—Zollweg and Gottlieb. (See 3122.)

Through the sophisticated use of hydraulics and air, IBM engineers and scientists have developed the IBM 1301 disk storage. In this new disk memory, a hydraulic actuator simultaneously positions 40 magnetic data heads to read and write on magnetic disk surfaces. The special design of the hydraulic actuator combines speed, accuracy and reliability. The magnetic heads glide on a self-acting air bearing which requires no external air supply. Heads float at precise distances above disk surfaces, supported only by air flow generated by the rotating disks.

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Industrial Engineering Notes*

Governmental and Legislative

President Kennedy last week approved the recommendations of his Space Council that communications satellites should be owned and operated by private industry. In a policy statement, the Chief Executive set forth these requirements for participants in the world-wide communications venture:

1. New and expanded international communications service must be made available at the earliest possible date;
2. The system must be global in coverage so as to provide efficient communication service throughout the whole world as soon as technically feasible, including service where individual portions of the coverage are not profitable;
3. Opportunities must be provided for foreign participation through ownership or otherwise, in the communications satellite system;
4. There must be non-discriminatory use of and equitable access to the system by present and future authorized communications carriers;
5. Effective competition, such as competitive bidding, must exist in the acquisition of equipment used in the system;
6. Structure of ownership or control must assure maximum possible competition;
7. Antitrust legislation with the regulatory controls of the government must be complied with; and
8. The system must be economical, the benefits of which will be reflected in overseas communication rates.

The government's role in the project, the President said, includes research and development, supervision of international agreements and negotiations, control of spacecraft launching, effective use of the radio spectrum, technical assistance to newly-developing countries, and examination, together with other nations, of how the United Nations can participate in international space communications. The Federal Communications Commission, in a simultaneous announcement, disclosed it had asked international communications carriers to organize an ad hoc committee to speed plans for their joint development, construction, ownership, and operation of a commercial satellite communications system. The Commission gave "prime importance" to quick development of the system and directed the committee to report by October 15. The committee was advised to seek aid and advice from representatives of industry and government agencies. Members of the ad hoc group are representative of the American Cable and Radio Corp. (Mackay Radio, Inc., The Commercial Cable Co., All America Cables and Radio, Inc., and Globe Wireless, Ltd.); American Telephone and Telegraph Co.; Hawaiian Telephone Co.; Press Wireless, Inc.; Radio Corp. of Puerto Rico; RCA Communications, Inc.; South Puerto Rico Sugar Co.; Tropical Radio Telegraph Co.; United States-Liberia Radio Corp.; and Western Union Telegraph Co.

The FCC also issued an order turning down a petition by the General Electric Co. and its subsidiary, Communication Satellites, Inc., to permit firms in the communication equipment and aerospace industries to share ownership of the satellite system. The Commission, in its First Report on the venture, said ownership should be restricted to international common carriers. The way was left open, however, for further action by General Electric. The FCC said it is still gathering preliminary information and that "any proposal or information that may evolve from this further stage of the inquiry, together with all information and proposals supplied by petitioners . . . and others, will provide a basis upon which the Commission, consistent with the public interest and applicable legal procedures, will take such further action . . . to achieve the expeditious establishment of a commercial satellite communications system." . . . **Electronic computer manufacturers last week were asked to propose plans and submit bids for a system to help process the thousands of applications the Federal Communications Commission receives from industry and the public.** Manufacturers were asked to base their plans on 440 pages of FCC-compiled electronic computer system specifications and make proposals by September 29. A proposed electronic computer system has been under study at the FCC since 1958. The FCC gave these examples of the jobs the computer might undertake: Compute AM, FM, and TV broadcasting radiation patterns; compile details about ownership of existing or proposed broadcasting stations; schedule pending radio license applications and renewals; help determine whether proposed antenna towers will create air safety hazards or cause broadcasting interference; analyze data to establish telephone plant depreciation rate values as a means to determine the validity of interstate telephone rates; and analyze frequency usage. . . . **A memory system installed in the TRANSIT IV-A has received information relayed to it from an injection station at Forest Grove, Md., and the satellite has sent the message back to two stations at the Applied Physics Laboratory of the Johns Hopkins University**

(Continued on page 102A)

* The data on which these NOTES are based were selected by permission from *Weekly Reports*, issues of July 31, August 7, 14 and 21, 1961, published by the Electronic Industries Association, whose helpfulness is gratefully acknowledged.

Discussing a system for space target detection (left to right): Dr. R. W. Bickmore, Technical Assistant to the President; Laboratory managers H. E. Shanks and M. G. Chernin; and M. D. Adcock, Director of the Electromagnetic Systems Division.

A Report from American Systems Incorporated...

Multiple Space-Target Detection

One of the key tasks of modern sensor technology is the determination of the positions and velocities of aerospace vehicles. A particularly important aspect of this problem, now being studied by American Systems Incorporated, is the radar detection of multiple targets. Usually such targets move in relatively small clusters, at high altitude and extremely high velocities. To track the individual objects making up a cluster, new antenna array techniques are under consideration.

This study is being conducted by the Electromagnetic Systems Division, the activities of which encompass research in electromagnetic physics, development of complete sensor systems, and design of special microwave components. Among the scientists and engineers contributing to this work are Dr. R. W. Bickmore and H. E. Shanks, inventors of Time Domain Radiation, and M. G. Chernin, microwave specialist. With active contracts in a number of important areas, the Division is contributing both to basic research, and to the creation of special microwave systems.

The Division's programs include Time Domain and Doppler approaches to measurement of the radiation patterns of large antenna systems, investigation of signal interference problems in areas of high density radio and radar installations, and the development of passive reflectors providing coded target returns.

Front line technical efforts are also under way in six other Divisions of the Company.

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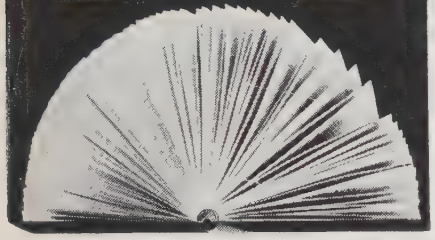

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Industrial Engineering Notes

(Continued from page 100A)

"with most satisfactory results," it was reported last week. The orbiting satellite is demonstrating the technique of time synchronization over large distances. TRANSIT time is based on a standard maintained by the U. S. Naval Observatory. The mem ory is a simple mechanical type which releases signals at precise intervals, measured by the extremely accurate pulses of an ultra-stable crystal oscillator, the announcement said. Transmission from the memory system to the ground takes about a half-minute and is immediately repeated. New data may be inserted in the satellite when the TRANSIT-IV-A approaches within line-of-sight of the injection station. Return information is being received from the satellite by two special stations at the Applied Physics Laboratory, The Johns Hopkins University, in Silver Spring and Howard County, Md. . . . **The 1961 edition of "Statistical Abstract of the United States" has been published by the U. S. Census Bureau.** Statistics for the new edition were furnished by 94 government agencies and 71 private and research organizations. It contains tables, charts, a bibliography, and an alphabetical subject index. Included in the 34 chapters are data on: Population; vital statistics, health, and nutrition; labor force, employment and earnings; income, expenditures, and wealth; prices; banking and finance; business enterprise; communications; power and scientific development; roads and motor vehicles; transportation, land and air; waterways, water traffic, and shipping; forests and forest products; fisheries; construction and housing; manufacturers; distribution and services; and comparat ve international statistics. The publication is available at $3.50 per copy from the Government Printing Office, Washington 25, D. C., or from any of the field service offices of the Department of Commerce in major cities around the country.

Contracts Awarded

Development of a compact airborne radar beacon for light aircraft to provide air traffic controllers with altitude information has been initiated by the Federal Aviation Agency, it was announced recently. Bids for a contract to develop a Small Lightweight Altitude Transmitting Equipment (SLATE) for use in high density terminal areas are being requested by FAA. The new equipment, which will consist of an altitude transducer and transponder system, must be capable of automatically reporting aircraft altitude in 500-foot increments from at least 10,000 feet down to sea level. Its range will be up to 50 miles. Under the proposal, the radar beacon will consist of a small, lightweight, inexpensive transmitter for the aircraft. When interrogated by a ground radar station the airborne equipment will return a pulse of energy. Altitude information will be relayed from the craft to the controller in the return transmission. The main purpose of the new equipment is to economically provide the user with the advantages of a safe, reliable positive separation service. The service is dependent upon altitude identification, in addition to range and azimuth information. . . . **The Martin Company recently received a contract from the Army to design, construct, and test a floating nuclear power plant to supply electricity to military installations or which could provide emergency power to port cities cut off from normal service by peacetime disaster or wartime action, it was reported last week.** The 10,000 kilowatt plant will be installed in the hull of a reconditioned and modified surplus Liberty ship. The total contract amounts to $16,998,127 but the actual award under it at this time will be limited to the design phase in the amount of $644,046. Construction, fabrication, and test operation phases are subject to the exercise of options by the government after the design has been completed. The contract provides that the design phase of the plant will be completed within 15 months. Subsequent construction and test phases will require an additional three years, it was explained.

Industry Marketing Data

Factory sales of transistors increased about $2 million and total value of units sold about $1 million in June, according to the latest figures released by the EIA Marketing Data Department. A total of 17,899,005 transistors worth $26,148,746

(Continued on page 104A)

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Professional Group on Circuit Theory

One of the most fundamental subjects in the radio engineering profession is circuit theory. It forms a basic part of every engineer's schooling. The method of analyzing the characteristics of circuits and the techniques of designing circuits for optimum performance find wide application in practically every branch of communications and electronics. It is not surprising, therefore, that a substantial portion of the technical literature is devoted to this important subject.

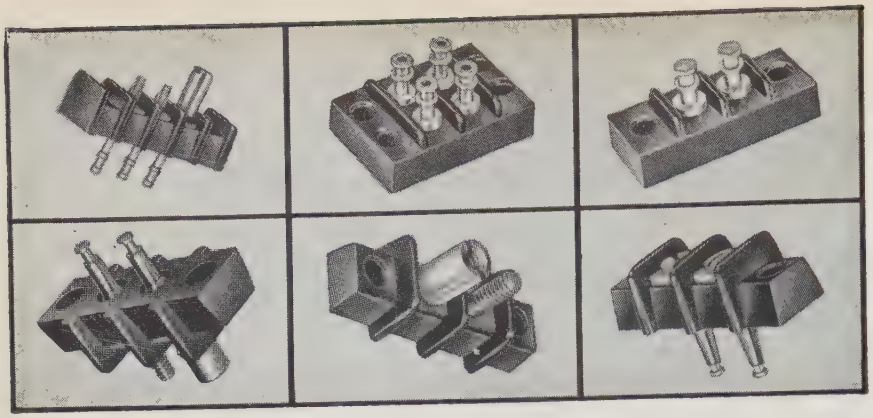
While circuit theory is broad in its applications, the formulation of these methods and techniques is a sharply defined and well-developed field in its own right, requiring specialized knowledge of a high degree. The IRE Professional Group on Circuit Theory is playing a key role in furthering the development of this field by making this specialized knowledge readily available to those who need it.

The Circuit Theory Group began operating in April, 1949 and since that time has become one of the largest and most active Groups in the IRE. Its most important activity has been the publication of its own technical periodical, called IRE TRANSACTIONS on Circuit Theory. Begun in 1952, the TRANSACTIONS is issued quarterly to its several thousand members as a part of their assessment. An interesting and valuable feature of TRANSACTIONS is that many of the issues are built around a special topic. Among the topics thus treated are Servo-mechanisms, Circuit Stability, Network Approximation, Nonlinear Filters, and Time-Variable Networks.

The Circuit Theory Group has been active also in sponsoring meetings and in organizing sessions at several national conferences. Numerous meetings are also held in over a dozen cities by local chapters of the Group. The net result has been to give engineers a golden opportunity to keep fully informed of the many significant developments in this important field.

Ernst Weber

Chairman, Professional Groups Committee



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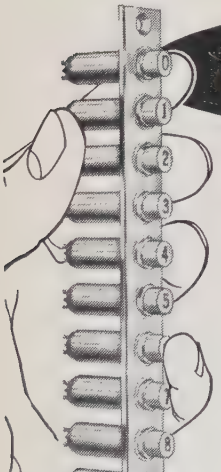

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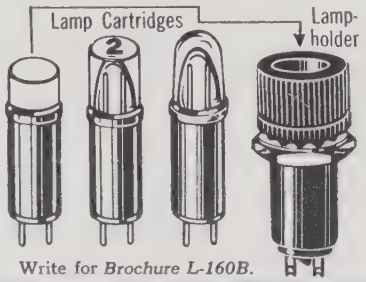

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Industrial Engineering Notes

(Continued from page 102A)

were sold at the factory last month. During the previous month, 15,128,181 units valued at $25,113,042 were sold. Through June of this year, sales totaled 88,682,882 transistors. Accrued revenue stood $157,-120,149. During the same period in 1960, 60,485,683 units were sold at a cumulative value of $152,932,961.

Section Meetings

ALAMOGORDO-HOLLOMAN

"Oscilloscope Sampling Concepts," Dean Butts, Tektronix, Inc. 6/29/61.

CENTRAL FLORIDA

"Improved Demodulation & Pre-detection Techniques," L. R. Rauch, Univ. of Mich.; "Review of JETS Program & Future Plans," Karsten Mikalsen, Kip Heintzelman, Melbourne High School. Joint with PGSET. 5/11/61.

"Description of N.A.S.A.'s Nimbus Satellite," Robert R. Golden, Nat'l Aeronautics & Space Administration. 6/15/61.

ELMIRA-CORNING

"Thermoelectricity." 3/27/61.

"Recent Advances in Materials: Thermionic Materials," Ivor Brodie, Westinghouse Elec. Co. 4/17/61.

"Recent Advances in Materials: Semiconductor Materials," Coleman Goldberg, Westinghouse Elec. Co. 4/24/61.

"Recent Advances in Materials: Metals," N. J. Grant, Mass. Inst. of Tech. 5/1/61.

"Recent Advances in Materials: Ceramics," Hans Thurnauer, Minn. Mining & Mfg. Co. 5/8/61.

Ladies Night—Dinner Dance. 5/24/61.

EL PASO

"Evolution of Sound Reproduction Equipment," F. L. Willems, The Howell Co. 7/27/61.

ERIE

"Structure of the Universe," C. M. Wheeler, GE Co. 6/27/61.

EVANSVILLE-OWENSBORO

"Rendezvous in Space," Anatole Browde, McDonnell Aircraft Corp. 2/15/61.

"Events in the Development of Receiving Tubes," W. T. Millis, GE. 3/15/61.

Tour of Alcoa Warrick Works Aluminum Plant. 4/12/61.

"Resonance (reverberation)," A. F. Petrie, GE; Annual Section Dinner. 5/10/61.

Annual Section Picnic. 6/18/61.

GENEVA

"Activities of the Geneva Section & International Plans of IRE," R. L. McFarlan, IRE Past President. 7/6/61.

"Future of Electronics," IRE International Activities Committee, R. L. McFarlan, Finley Carter, Dr. Henderson, Ir. H. Rinia, E. Weber. 7/7/61.

(Continued on page 106A)

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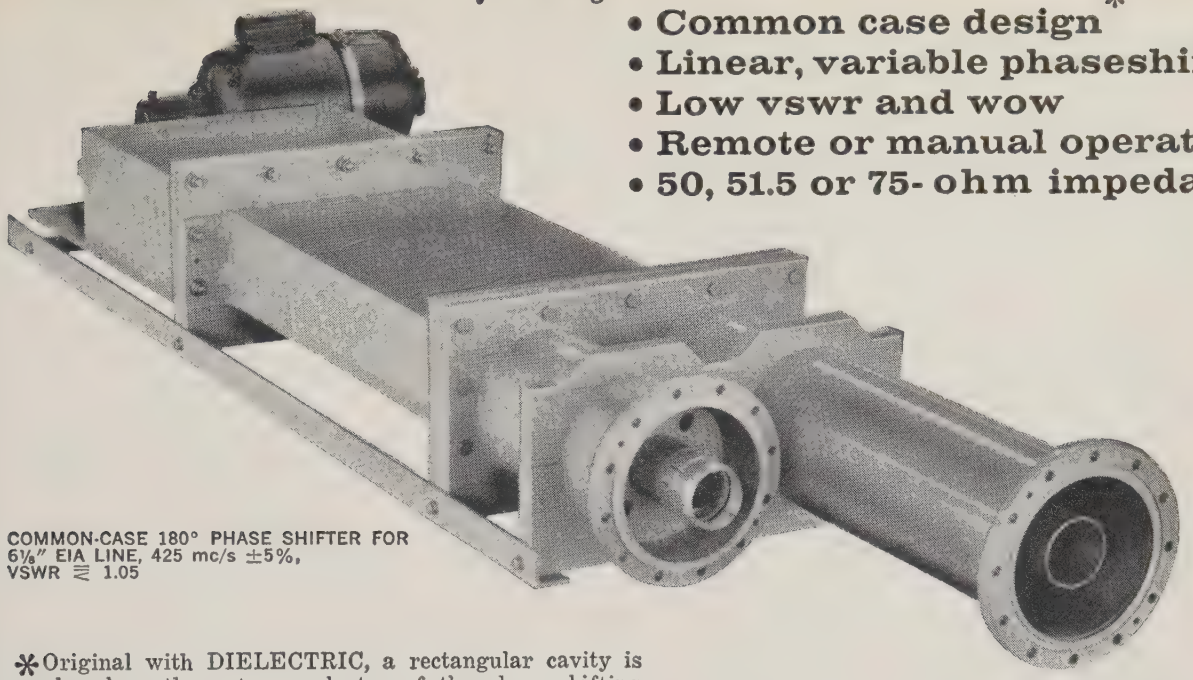

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In gaining these several advantages, common-case phase shifters do become limited in band width by the VSWR specification. Their usual application is in systems where the operating frequency is within a 20% band, ±10% about a specified center frequency. But for every application within this one limitation, common-case phase shifters will satisfy every performance need.

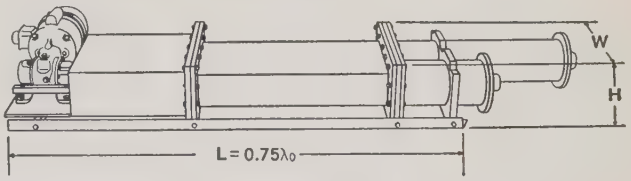
The phase shift is very linear because of the simplified mechanical structure, and is continuously variable. Both manually and electrically controlled models are available.

In the latter case, adjustable electrical limit stops as well as positive mechanical stops are provided. For remote, electrical operation, a precision potentiometer is connected directly to the travel mechanism for accurate position information. These units can be furnished in any line size from ⅞" OD and up; 50-, 51.5-, or 75-ohm impedance; with either EIA (RETMA) or QUICK-CLAMP flanges, or without flanges.

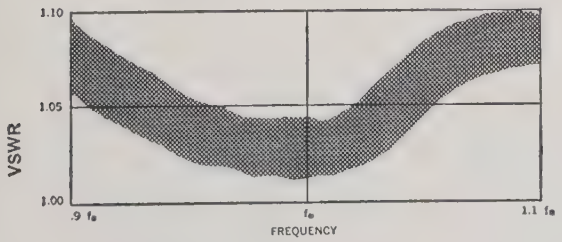
Power handling capacity, upper frequency limits, and approximate physical dimensions are shown in the accompanying table and outline drawing. Also illustrated is typical graph of VSWR vs Operating Frequency. The vertical dimension of the grey band is the VSWR variation over 180° of phase shift.

To order, specify frequency range, line size, line impedance, flange type, phase shift/minute if motorized, and position-indicator details. For further information including price, contact our Proposal Engineering Department.

OUTLINE DRAWING — 12000 SERIES



TYPICAL VSWR vs OPERATING FREQUENCY
(Vertical height of band is variation over 180° of phase shift)



CONDENSED SPECIFICATIONS — 12000 SERIES

LINE SIZE (in.)	POWER (kw, max)	UPPER FREQUENCY LIMIT	APPROXIMATE DIMENSIONS (inches) W	H	L
⅞	12 @ 1, 4 @ 100, 1 @ 1000 mc/s	3600	5	4	0.75λ0
1⅝	15 @ 100, 4.5 @ 1000, 2.4 @ 3000 mc/s	3200	6	4	
3⅛	150 @ 10, 48 @ 1000 mc/s	1000	10	5	
6⅛	600 @ 10, 190 @ 100 mc/s	500	14	7	
9 3/16	1400 @ 10, 380 @ 100 mc/s	350	24	12	0.75λ0

DIELECTRIC'S areas of capability include coaxial, waveguide and open wire techniques . . .

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(3) Attending AFCEA chapter meetings, dinners and a big annual exposition for publicizing your firm and displaying your products.

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BPA

Section Meetings

(Continued from page 104A)

HAWAII

Annual Social/Business Meeting; Election of Officers. 6/16/61.

HOUSTON

Annual June Picnic. 6/16/61.

LAS VEGAS

"First & Second Generation Space Probes," A. K. Thiel, Space Technology Labs. 8/7/61.

NEW YORK

"Modern Trends in Graduate Engineering & Research"—Panel Discussion: H. J. Carlin, PIB, J. H. Mulligan, NYU, R. J. Schwarz, Columbia. 5/3/61.

NORTHERN NEW JERSEY

Visits to Electronics Association Computation Center and Plasma Physics Lab. 6/15/61.

OMAHA-LINCOLN

"Allocations, Coverage, Consternation—Where Is TV Signals Really Go?," A. J. Ebel, KOLN-TV; "KFAB Stereo—The Present AM-FM, & Future Multiplex FM," G. A. Totten, KFAB; Section's 13th Birthday Party. 8/4/61.

REGINA

"Equipment Used in Cancer Clinic," H. Crosby, Grey Nuns Hospital. 5/1/61.

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"Upper Atmospheric Physics," B. W. Currie, Univ. of Saskatchewan. 5/16/61.
"A Talk on the IRE—Past & Future," J. F. Byrne, Vice President of IRE. 6/15/61.

RIO DE JANEIRO

"The Problem of Measurement of Short Time Toll Signalling Pulses," D. S. Marinho, Companhia Telefonica Brasileria. 5/10/61.
"Single Side Band Transmission Systems," A. C. Jorge, AJ Electronica Ltda. 6/14/61.

ST. LOUIS

Student Paper Presentation. 5/17/61.
Business Meeting. 6/13/61.

SCHENECTADY

"Bionics," G. M. White, GE. 7/5/61

SUBSECTIONS

BURLINGTON

"Electron Tubes & Semi-conductors of Past, Present, & Future," R. E. Moe, Director of Region V. 5/25/61.

CATSKILL

"Better Engineering Writing," R. M. Koff, Product Engrg. Magazine. 2/28/61.
"Gravity, Inertia & The Universe," D. W. Sciama, Cambridge Univ. & Cornell Univ. 3/14/61.
Tour of SAC Test Facility. Joint meeting with PGEC N. Y. Chapter. 4/20/61.
Annual Dinner Dance; Election of Officers. 6/17/61.

PASADENA

"Current Trends in Engineering Education," F. C. Lindvall, Cal Tech. 6/13/61.

(Continued on page 108A)

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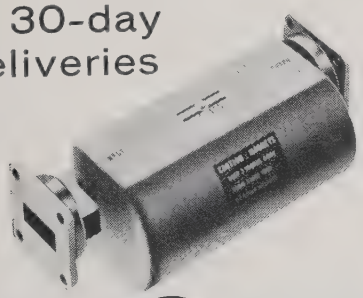

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Section Meetings

(Continued from page 107A)

RICHLAND

"Noise in Electron Tubes," Cris Metelmann. 3/15/61.

"Solid State Devices," R. Sherrard, M. T. Slind, G. R. McKinney, GE. 3/22/61.

"Adaptive Control Systems Research at the University of Washington," R. N. Clark, Univ. of Wash. 4/26/61.

"Some Recent Trends in Signal Analysis," H. L. Libby, GE. 5/24/61.

"The Mysteries of Color," Jim Russell, Byron Brenden, GE. 6/14/61.

SANTA BARBARA

"The Engineer as Entrepreneur"—Panel Discussion: H. W. Brelsford, Wm. Bollay, R. Redemske; Election of Officers. 5/16/61.

Dinner-Dance. 6/24/61.

VICTORIA

Inaugural Address, W. Thompson, Vancouver Section Chairman. 4/26/61.

"Underwater Sound Propagation," H. Davidson, Pacific Naval Lab. 5/24/61.

Positions Wanted

By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The IRE publishes free of charge notices of positions wanted by IRE members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The IRE necessarily reserves the right to decline any announcement without assignment of reason.

Address replies to box number indicated, c/o IRE, 1 East 79th St., New York 21, N.Y.

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Experienced teacher at graduate and undergraduate level in E.E. and physics. Ph.D. Specialist in planning of new programs. Desires position of responsibility where teaching, research, publications and organizational experience are fully utilized. Late 30's. East or west coast preferred. Box 3946 W.

ENGINEER—PILOT

BSEE. Member Tau Beta Pi. Age 32. 4000 hours pilot time including 1500 hours jet time. Presently Capt. USAF doing engineering/pilot duties. Experienced in airborne and ground navi-

(Continued on page 110A)

America's FIRST two-man space team will rely on a Honeywell designed and developed inertial navigation and flight control system for controlling their space craft. The work we are doing on Project Dyna Soar is representative of the many advanced systems that are being evolved and developed into working hardware at Honeywell. It is typical of the creative concepts that pose challenges to the engineer or scientist possessing imagination, drive and talent.

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Sr. Flight Systems Engineer—Experience in the systems rather than component design in stability and control of aircraft and/or missiles. Should be familiar with the techniques associated with satellite rendezvous, trajectory and re-entry analysis.

Sr. Systems Analyst—Experience in analysis of aircraft stability and control. Familiar with both analog and digital computers and capable of applying various types of stability criteria such as phase amptitude plots, root locus, etc.

Transistor Circuit Applications—Several years experience in audio frequency circuit design or servomechanisms circuits utilizing transistors, diodes, transformers, magnetic amplifiers, etc.

Microelectronic Engineer—Experience in the basic development of micro-miniature devices or design of one or more types of electronic components such as semiconductors or resistors.

Reliability Engineer—Determination and application of reliability techniques on components and systems. Knowledge of basic electronic circuits and reliability concepts such as prediction, derating, parameter variations and redundancy.

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ARMOUR RESEARCH FOUNDATION

Electromagnetic Compatibility Analysis Center, NEES Annapolis, Maryland

By Armed Forces Veterans

(Continued from page 108A)

gation and communications equipment. Considerable management and supervisory experience. Interested in a challenging position which utilizes both engineering and pilot experience. Box 3947 W.

SUPERVISORY OR MANAGERIAL

Supervisory or managerial type position is desired; B.S. in mathematics and physics; M.A. in administration and mathematics; 3 years industrial experience as supervisor of data reduction section; 6 years teaching experience, 4 at the college level and 2 in high school; 4 years experience as Executive Officer USNR; age 34; married with a family. Box 3948 W.

SERVO ENGINEER

BSEE. 1959, E.I.T. 1959; 7 months missile experience; 1½ years R&D design and checkout of fire control equipment. Additional graduate work in engineering and business. 7 months Naval Electronic School; 2 years as Naval Electronics Technician. Desires position in design or sales engineering with opportunity for advancement and further study. Age 27; married, 1 child. Willing to relocate. Box 3949 W.

COMPUTER ENGINEER

BEE., R.P.I. 1956, MSEE., M.I.T. 1958, MS. Physics, N.U. 1961. 5 years experience in computer systems and component development. Looking for challenging position in computer development, applications and/or sales. Desires New York City area. Willing to travel. Box 3951 W.

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Electrical Engineer desires challenging position in R&D work requiring professional competence. Experience in wire communications, data processing and transmission, programming, microwave systems, teaching, technical writing and supervision. Prefer area where graduate study is available. Will forward detailed resume on request. Box 3959 W.

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ENGINEER—MANAGER

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(Continued on page 112A)

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It's a big job. To do it right, JPL must have the best technical people in the country. People who want to know . . . who want to be part of the greatest experiment of mankind. If you're that kind of people, JPL is your kind of place. Write us today.

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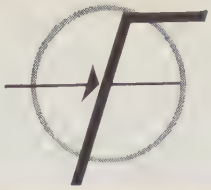

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Positions Wanted

By Armed Forces Veterans

(Continued from page 110A)

military electronics systems, including ECM, aircraft C & N, and radar. Eight years of large corporation management experience. Prefer suburban area in Mid-Atlantic region. Box 3964 W.

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B.S.E.E., 8 yrs. experience in Electronic Component sales plus two years experience as Project Engineer for Component Manufacturer. Successful record in increasing sales. Tripled sales in present territory in three years. Box 3965 W.

Positions Open

The following positions of interest to IRE members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. ....

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Seeking experienced, qualified television technician for maintenance and operation at commercial VHF station in growing market. CBS affiliate; progressive management with long broadcast experience. Station is well-equipped; maximum power; 1000 foot tower; two Ampex Videotape Recorders. Applicant must be ambitious, dependable, and have 1st class phone license. Some formal schooling in electronics preferred. Salary commensurate with experience and ability. Replies treated in confidence. Write Chief Engineer, WLAC-TV, Nashville, Tenn.

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(Continued on page 116A)

GM's DSD tracks down the answer!

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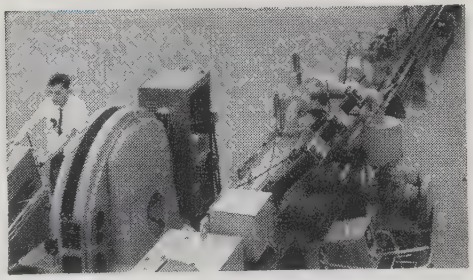

Two Light Gas Guns permit flight of hyper-velocity projectiles . . . including studies of high-speed impact, properties of ionized gases, new methods of detection and identification. If you can use facilities like these, contact DSD.

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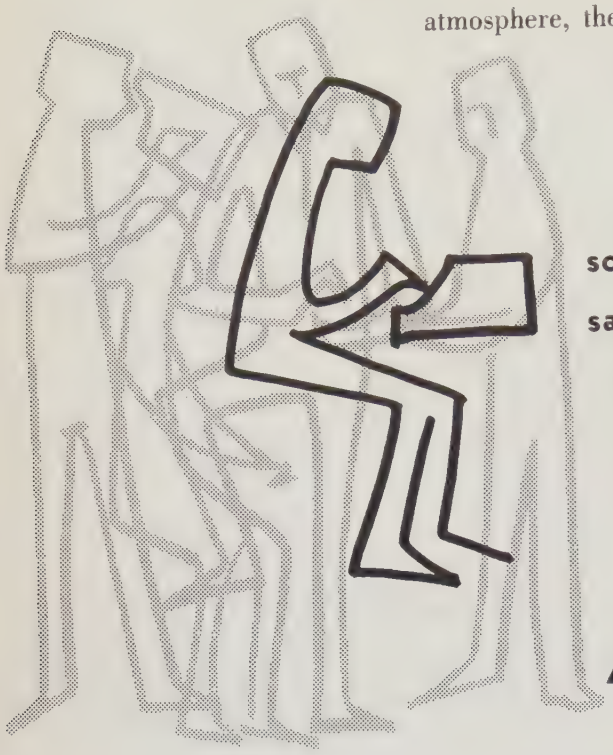
Lenkurt Electric's engineers create unique telecommunications systems for industrial, telephone, and defense requirements. Working in an independent and creative atmosphere, they have every opportunity to further their formal education, live in the West's outstanding climate, and enjoy unlimited cultural pursuits. Qualified *communications engineers* are now being added to the staff in the specialized fields of:



solid state circuitry • data transmission

sales • mobile radio • data transmission

Qualified specialists in these fields should send a resume or call collect to: *E. Jack Shannahan, Employment Manager • Lenkurt Electric Co., Inc. • 1105 County Road • San Carlos, California. LY 1-8461, Ext 281*

LENKURT ELECTRIC

Subsidiary of

GENERAL TELEPHONE & ELECTRONICS

GENERAL SYSTEM

ABBOTT'S FINEST CAREER OPPORTUNITIES FOR OCTOBER

Through close association with the nation's leading employers of scientists and engineers, Abbott's makes available to you exceptional employment opportunities. The following positions have been selected as our best career opportunities this month, because they offer unusual advancement and creative potential, in addition to excellent salaries and fringe benefits.

SALARY RANGES: $10,000-$25,000

SCIENTISTS—Ph.D. E.E.—Solid State Physics—Advanced Mathematics

PHYSICISTS—Senior Development—Theoretical—Astro-Nuclear Creative R. & D.

ELECTRONIC ENGINEERS—All levels—Pulse Circuitry—Digital Systems—Analysis Design & Packaging

MANAGERS—Microwave Components and Systems—Antenna Research and Navigation—Servo Mechanisms—Electro Hydraulic and Pneumatic

PROGRAMMERS and MATHEMATICIANS

Abbott's placement service is available at no charge to you. All expenses and service charges are paid by our clients. All negotiations conducted in strict confidence.

MR. LOUIS A. KAY

ABBOTT'S

EMPLOYMENT SPECIALISTS

150 Tremont Street, Boston 11, Massachusetts
HAncock 6-8400

Positions Open

(Continued from page 112A)

ANTARCTIC RESEARCHER

Douglas Aircraft has an opening at Antarctic station for experimental investigations of the ionospheric absorption of extraterrestrial radio waves. Applicants must have an advanced degree (or equivalent) in physics or electrical engineering with a knowledge of ionospheric physics and/or solar flare events and r. f. propagation in the atmosphere. Experience required in data reduction and analysis, scientific instrumentation, and communication systems. Applicants should also be familiar with operation of a riometer, although not required. In order to qualify, individuals must be 21 years of age, single (preferred), and will be required to pass stringent medical examination. Apply to S. Amestoy, Douglas Aircraft Co., Inc., 3000 Ocean Park Blvd., Santa Monica, California.

OVERSEAS FIELD ENGINEER

Electronic Field Engineers are needed for overseas assignments. Applicants must have a B.S. degree and at least 1 year of experience in electronics and must possess initiative, ingenuity, and the ability to get a job done. Base salary during training in Denver with overseas differential and per diem while on overseas assignment. Address inquiries to: C. A. Hedberg, Head, Electronics Div., Denver Research Institute, University of Denver, Denver 10, Colorado.

ANTENNA ENGINEER

Experienced in antenna design and project work in the field of communications antenna systems. Write including complete resume and salary requirements to Roger Olson, Director of Engineering, Hy-Gain Antenna Products Corp., 1135 North 22nd St., Lincoln, Nebraska.

DATA PROCESSING ENGINEER

An attractive position is open for an engineer or mathematician with one to four years experience (or the equivalent in graduate education) who would like to attack challenging problems such as the analysis of automatic control, data processing, and tracking systems. Education and/or experience in disciplines such as control systems, statistical communications theory, digital systems, and network analysis desirable. Opportunity to work with an energetic group of talented engineers in a conducive research atmosphere. Contact Mr. A. P. Rentschler, Employment Mgr., Cornell Aeronautical Lab., 4455 Genesee St., Buffalo 21, New York.

ENGINEERS

Two Electrical Engineering positions open with Application Engineering Dept. charged with the responsibility of assisting the technical groups of customers and prospective customers in the application of products manufactured by the company, with especial emphasis on the application of capacitors to equipment and systems. This group supplements the work of the sales personnel in the field, works closely with the Design and Product Depts. One Engineer desired with strong background in electronic circuitry and filter design; second Engineer, with experience with electronic computers and computer circuits. Location: East coast. Write including personal resume and curriculum vitae. Box 2059.

(Continued on page 118A)

AWCS-412-L Consisting of a closely coordinated network of data acquisition stations, data processing and display centers and weapon bases, **Air Weapons Control System 412-L** provides the tools for effective and flexible air space management, continent-wide or in single point defense. Vital detection and tracking information is supplied automatically to human decision-makers within seconds. Effective direction of both manned and unmanned weapons, including return of manned aircraft to base, is a system function.

412-L STAFF EXPANDING

as program moves ahead in advanced development phase

If you are one of the fast-growing group of scientists and engineers who recognizes the unequaled career advantages and satisfactions to be gained through work on very large-scale, complex systems, we urge you to scan this abbreviated list of current openings on 412-L. If your background, experience or interests are represented, write us today. We promise you a careful review of your qualifications and a prompt reply.

COMMUNICATION SYSTEMS

Training in information theory and switching theory; broad experience in both RF and wire communication systems design, especially automatic subscriber and dial systems; to determine and specify characteristics for intra- and inter-site command and control complexes.

OPERATIONS ANALYSIS

Involves synthesis/analysis of computer oriented weapons control systems (ground/air environments) from an operational standpoint. Estimates system (personnel & equipment) capabilities; derives and evaluates system procedures using analyses and computer simulation as tools. BS/MS – minimum 3 years experience.

EQUIPMENT EVALUATION

Solving man-machine problems, evaluating alternative components, displays, or techniques, devise simulators. Advanced degree in experimental psychology required.

An Equal Opportunity Employer

SYSTEM EQUIPMENT ANALYSIS

Performs analysis on system-equipments and their functions; derives criteria and parameters to meet operational requirements. Work in areas of signal processing, detection, tracking and digital communications. Requires BS/MS/EE/Physics/Math—knowledge of simulation techniques, feedback theory, probability and information coding desirable. 3 years experience.

WEAPONS INTEGRATION

Determines and specifies optimum weapon utilization. Specifies weapon interfaces to equipment designers to insure system compatibility. Support / provide analytical inputs relative to weapons capabilities; analytical studies to assure optimum weapon employment within specific computer capabilities. BS/MS. 2 years manned/unmanned weapons system experience.

INFORMATION PROCESSING & DISPLAY SYSTEMS

Training in information and statistical theory; broad experience in design and integration of tactical and strategic data processing systems; to define system parameters, identify interface problems, perform math analyses of control loops as applied to real-time processing and display systems.

TECHNICAL WRITERS

Creation of systems manuals, engineering reports, proposals; required is at least 2 years of technical education with minimum of 5 years in engineering writing. Knowledge of graphic arts techniques would be helpful.

PROGRAMMING ANALYSIS

Degree required. Math background; 2 years experience on medium/large-scale digital computers solving real-time military operational problems. Activity ranges from problem analysis to program development, including flow diagramming, coding and debugging in the utility control, simulation and operation areas.

ANTENNA & MICROWAVE COMPONENTS

BSEE, advanced degree preferred; experience with high power phenomenon, antenna development, RF plumbing, modulators, transmitters and receivers.

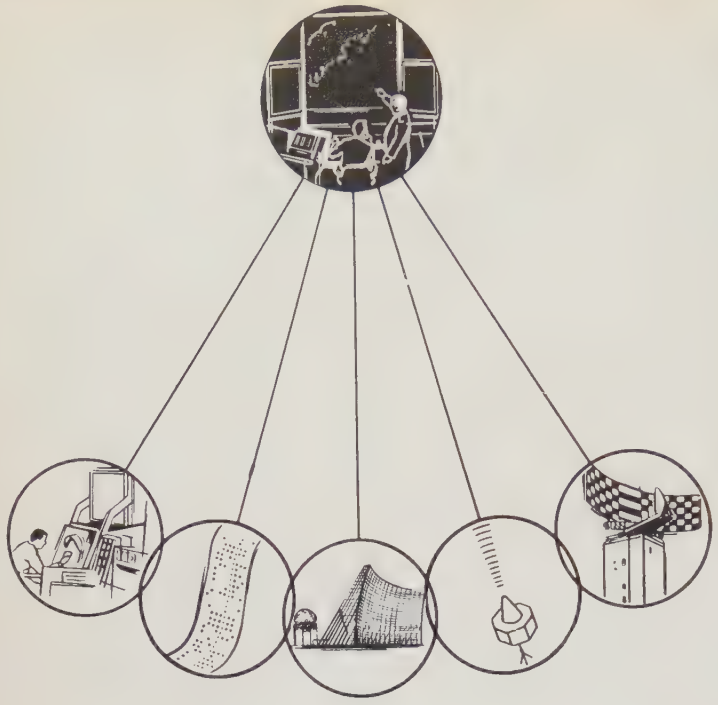
APPLIED MATHEMATICS

Performs mathematical analyses on all phases of ground electronic control systems. Utilizes probability theory, numerical analysis, complex variables; knowledge and utilization of analog/digital computers desirable. BS/MS Math required; 2 years experience on applied math problems in engineering field.

Write in confidence to Mr. P. W. Christos, Div. 53-MJ

DEFENSE SYSTEMS DEPARTMENT (A Department of the Defense Electronics Division),
General Electric, Northern Lights Office Building—Syracuse, New York

GENERAL ELECTRIC



COMPRESSED TIME...AND A NATION'S NEED

The time available between alarm and decision has shrunk to minutes. In this compressed time, pertinent information must be gathered, transmitted, evaluated and displayed to the commanders through a variety of systems.

Design, development and evaluation of such command and control systems for the urgent present and the uncertain future is the vital function of MITRE.

Systems such as SAGE, BMEWS, MIDAS, Strategic Air Command and Control System, NORAD Combat Operations Center and others are all within the scope of MITRE'S system integration work for the Air Force Electronic Systems Division.

The job is challenging — the opportunity exists to break out of a single specialty — the reward is not confined to the financial.

Engineers and scientists interested in the vital field of command and control technology are invited to inquire about openings in:

- SYSTEM ANALYSIS
- OPERATIONS RESEARCH
- COMMUNICATIONS
- ADVANCED SYSTEM DESIGN
- ECONOMICS
- MATHEMATICS
- ECONOMETRICS
- COMPUTER TECHNOLOGY
- HUMAN FACTORS
- RADAR SYSTEMS AND TECHNIQUES
- ANTENNA DESIGN — MICROWAVE COMPONENTS
- AIR TRAFFIC CONTROL SYSTEM DEVELOPMENT

Write in confidence to: Vice President — Technical Operations, The Mitre Corporation, P.O. Box 208, MX4, Bedford, Mass.

THE MITRE CORPORATION

All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin

Positions Open

(Continued from page 116A)

PROFESSORS

The Air Force Institute of Technology, Wright Patterson Air Force Base, Ohio, has a vacancy in the Department of Mathematics. Most of the work is at advanced undergraduate and graduate level. One quarter out of four free of teaching duties for research or other academic pursuits, plus vacation. Working conditions comparable to those in leading universities. Employment will be effected in accordance with Civil Service regulations. Rank and salary will depend upon qualifications of applicant and may be anywhere between assistant professor, GS-11, $7,560.00 and full professor, GS-14, $12,210.00. Applications should be made on Standard Form 57, available at any Post Office or by letter to Head of the Department of Mathematics, Institute of Technology, Wright Patterson AF Base, Ohio.

GRADUATE ASSISTANTSHIPS

Research and Graduate Assistantships in Instrumental Sciences leading to a M.S. degree in Instrumental Sciences. Engineering and Physics majors with strong background in instrumentation. Course work in fundamental principles and theory of basic instrumentation and process control. Research work in many fields such as high vacuum, electron and ion generation, mass spectrometry and gas chromatography, and process instrumentation. Stipend $2100 to $2400 for twelve months. Send transcript and vitae to: Head, Electronics and Instrumentation, University of Arkansas, Graduate Institute of Technology, P.O. Box 3017, Little Rock, Arkansas.

ENGINEERS MATHEMATICIANS PHYSICISTS

Engineering-Physics Co. has recently embarked upon new fields of research and development. Singular opportunities exist for engineers and scientists who have the resourcefulness, imagination and technical background to assume a responsible role. All qualified applicants will be considered for employment regardless of race, creed, color or national origin.

Projects include studies and developmental work on:

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MAGNETOHYDRODYNAMICS
ELECTROMECHANICAL DEVICES
SYSTEMS CONCEPT & DEVELOPMENT

Address inquiries to Dr. Vincent Cushing

EPCO ENGINEERING-PHYSICS COMPANY

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Opening for Ph.D. in physical chemistry, physics or metallurgy. We are producers of highest quality silicon and one of our chief research interests in this field is vapor-deposited single-crystal layers. We are interested in having as accurate and comprehensive information as possible on such structures and on their possible application to devices. We are already engaged in electrical and optical studies and are most desirous of deepening and broadening these studies. A sufficiently qualified applicant would have ample opportunity for choosing and exploiting directions as well as specific problems of interest. Other fields of research in semiconductors are gallium arsenide, germanium and thermoelectric materials. Reply in confidence to Box 2060.

SENIOR MATHEMATICIAN—STATISTICIAN

Supervisory position for a Senior Mathematician experienced in the use of Gaussian, Exponential, Gamma and Weibull Distributions for reliability predictions in electronic circuitry. Apply to Melpar, Inc., 3013 Arlington Blvd., Falls Church, Va.

MEDICAL ELECTRONICS

A rapidly expanding medical school requires an engineer to head an electronics development service. A graduate degree with an interest in the bio-medical field or a B.S. degree with considerable applicable experience is required. Contact Dr. John Schuder, University of Missouri, Medical Center, Columbia, Missouri.

COGNITIVE SYSTEMS SCIENTIST

A staff opening exists in the Computer Research Department of the Cornell Aeronautical Laboratory which will permit a resourceful, imaginative scientist to participate in research programs on cognitive, self-organizing and adaptive systems.

(Continued on page 120A)

ARIZONA
Phoenix

"Opportunity & Ideal Living"

GOODYEAR AIRCRAFT CORPORATION
Arizona Division

ELECTRONIC CIRCUITRY

Radar Circuitry, Data Processing, Analog Computer Circuitry, Control Systems.

THEORETICAL PHYSICIST

Background in Electromagnetic Theory and Plasma Physics, Ph.D. Preferred.

EXPERIMENTAL PHYSICIST

Background in Physical Optics, Classical Mechanics, Microwaves, and Information Theory or Quantum Mechanics and Statistical Mechanics, Ph.D. Preferred.

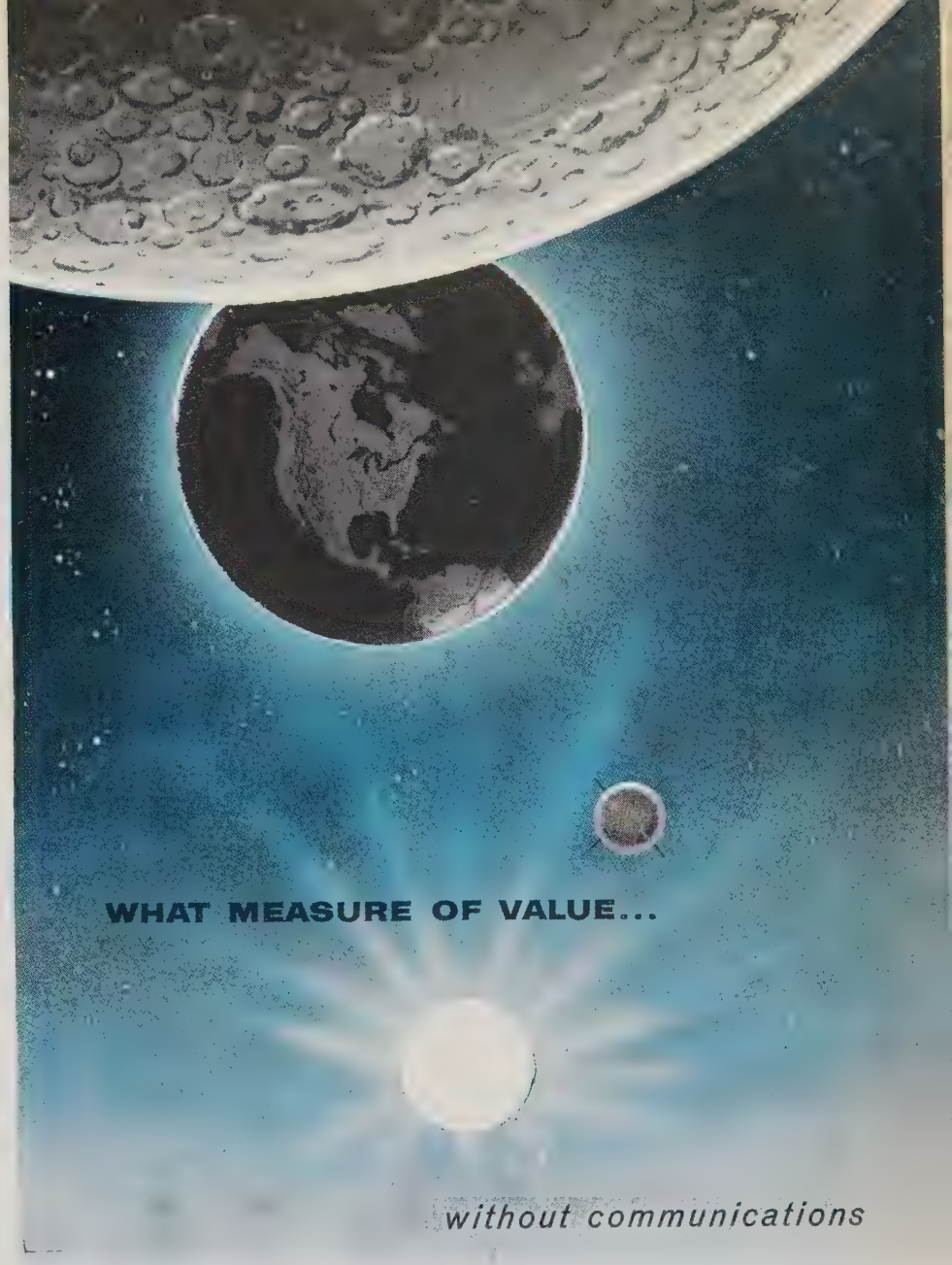
ELECTRONIC PACKAGING

"All qualified applicants will receive consideration for employment without regard to race, creed, color, or national origin."

Request Application or Send Resume to:

B. A. Watts, Engineering Personnel
Goodyear Aircraft Corp.
Litchfield Park, Arizona

Similar Positions at Goodyear Aircraft Corporation, Akron, Ohio



without communications

Satellites orbiting earth . . . satellites orbiting other planets . . . manned space stations. They represent the today and tomorrow of man's probing of our universe. Consider their value, however, were we not able to communicate with them . . . nor they with us.

At Amherst Laboratories state-of-art advances in microwave and millimeter wave technology are playing an important role in determining the future of aerospace communication and instrumentation systems. Significant advances in this field are being shaped today through individual and team efforts.

PROFESSIONAL STAFF AND MANAGEMENT OPPORTUNITIES are unlimited for Physicists, Mathematicians and Electronic Engineers with advanced degrees and creative desire. All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin. You are invited to direct inquiries in confidence to Mr. C. H. Haushalter, in care of Microwave and Aerospace Systems Department,

AMHERST LABORATORIES • 1181 WEHRLE DRIVE • WILLIAMSVILLE, NEW YORK

SYLVANIA ELECTRONIC SYSTEMS
Government Systems Management
for GENERAL TELEPHONE & ELECTRONICS

GENERAL SYSTEM

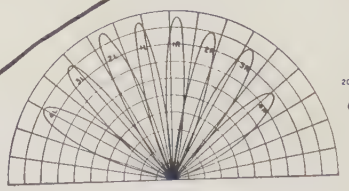
"Shrinking the Universe . . . through Communications"

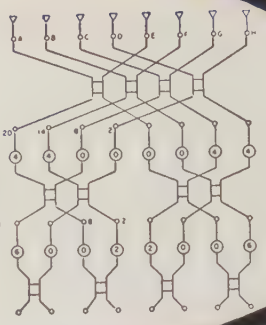
How Sanders

meets rising need for design simplification of sophisticated electronic equipments

New Beam-Forming Matrix for Electronic Scanning at Microwave Frequencies

This new Sanders development drastically reduces the astronomical number of components required by conventionally designed antennas. For instance, a 64 x 64 array requiring almost half a million 3-port dividers needs less than 25,000 hybrids with Sanders' new techniques. Accompanying diagrams illustrate radiation pattern and eight-element, eight beam matrix, formed by TRI-PLATE® components, a Sanders' first dating back to 1952.





Equally Challenging Assignments
Open to Engineers
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Pulse Doppler Radar Systems
Space and Communications Systems

Specific background required in
SYSTEMS DESIGN • TRANSISTOR CIRCUIT DESIGN
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This phenomenal growth company—up from 11 men to over 2000 in 10 years—today has a backlog of over $90 million in diversified R&D and production contracts. Concentration on bold technical innovations rather than marginal improvements is the dynamic force behind this growth. If you believe your career will flourish in a creative atmosphere that favors the individual contributor, send a resume now to R. W. McCarthy.

All qualified applicants will be considered without regard to race, creed, color or national origin

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Positions Open

(Continued from page 119A)

The several active and expanding programs now underway were developed from—but are not limited in scope by—the earlier conception and development of perceptrons at this not-for-profit research organization. Current analytical and experimental work includes pattern recognition problems such as those encountered in photo interpretation, target classification, and speech recognition.

In the past we have found that electrical engineers, mathematicians, and psychologists with training and/or experience in such areas as information theory, statistics, adaptive control systems, advanced programming, automata theory, brain models and learning mechanisms—frequently possess suitable capabilities and interests. We invite your application. Salaries are competitive with offerings of private industry, working conditions are excellent, and outstanding talent is appropriately rewarded. Contact A. P. Rentschler, Cornell Aeronautical Laboratory, Inc., 4455 Genesee Street, Buffalo 21, N.Y.

Use your
IRE DIRECTORY
It's valuable!

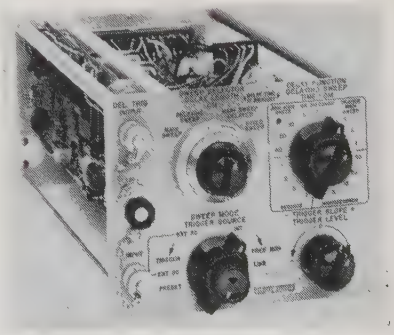
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 96A)

Plug-Ins Increase Oscilloscope Versatility

Two new horizontal plug-in units which increase the versatility of Model 160B and 170A oscilloscopes are now available from **Hewlett-Packard Co.**, 1501 Page Mill Rd., Palo Alto, Calif. The Model 160B is a 15 mc oscilloscope, while the 170A is a 30 mc unit. Both are built to military specifications.



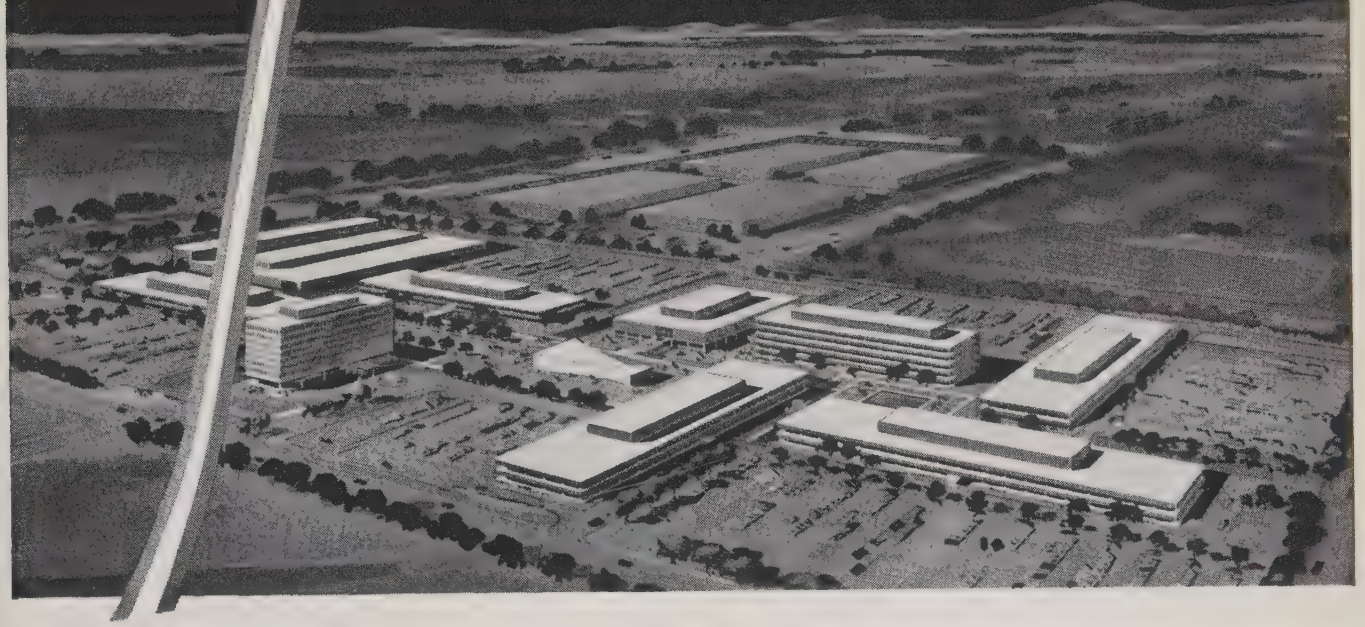
The new time-axis plug-ins are the Model 166C Display Scanner and the Model 166D Sweep Delay Generator.

(Continued on page 122A)

STL SPACE TECHNOLOGY CENTER

1 SPACE PARK

This November, eleven months after groundbreaking, the free world's most experienced industrial space scientists and engineers will be at work in STL Space Technology Center, on a 110-acre site at One Space Park, Redondo Beach, California. The spectacular compression of time, in conceiving, planning and constructing America's newest and largest facility for advancement of space science and technology, is testimony to the effectiveness of applying the systems approach to construction. It is made possible by pooling the experience, knowledge, management and technical skills of the owner, Space Technology Laboratories, Inc., the architect, Albert C. Martin and Associates, and the engineer-constructor, Twaits-Wittenberg Co.



STL Space Technology Center is a planned environment for Space Technology Leadership, and its construction is a milestone marking the continuing growth and diversification of Space Technology Laboratories, Inc. The outstanding facility of its kind, it is expressly designed for the outstanding men of their kind: engineers and scientists whose high order of achievement is demonstrated by their contributions to the nation's ballistic missile and space programs. The increasing requirements of STL's responsibilities in such programs as OGO, ADVENT, ATLAS, TITAN, and MINUTEMAN, create immediate opportunities in Southern California and at Cape Canaveral for engineers and scientists of talent and vision. All qualified applicants, regardless of race, creed, color or national origin, are invited to communicate with Dr. R. C. Potter, Manager of Professional Placement and Development.

SPACE TECHNOLOGY LABORATORIES, INC. P. O. BOX 95005-1 LOS ANGELES 45, CALIFORNIA

a subsidiary of Thompson Ramo Wooldridge Inc.

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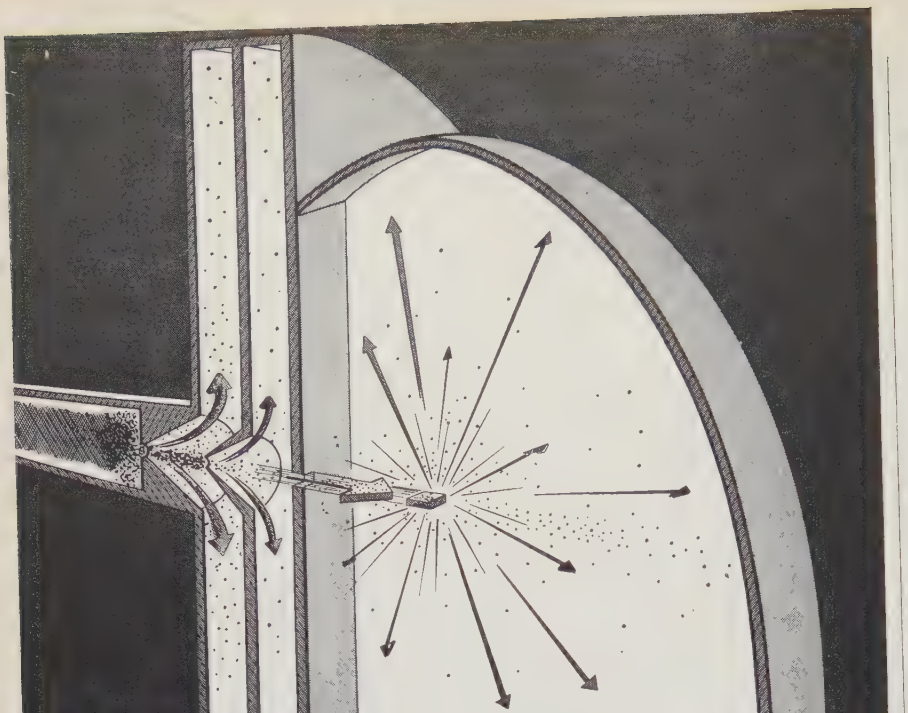
Those attending ARS Space Flight Report are invited to visit STL Booth #216



GAS PROPERTY STUDIES
BY MOLECULAR BEAM

• Depicted above is a tailored interface hypersonic shock tunnel used as a source of high speed neutral atoms and molecules for a "molecular" beam. With this beam, the constituents of air are made to collide at the relative kinetic energies appropriate to the high temperatures encountered in hypersonic flight. Too hot to be produced by an oven, yet too slow to be obtained conveniently from an ion beam by charge exchange, these particle energies can readily be obtained from a shock tunnel. In the hypersonic nozzle random energy is converted into well-directed uniform translational motion. When scattered from a gaseous target, this beam will allow differential cross sections to be measured and provide much needed data on which to base better calculations of gas properties. When scattered from a solid surface, the beam can provide information about the exchange of energy between the beam and the surface as well as information about the structure and chemical characteristics of the surface.

At CAL we are engaged in a wide variety of fundamental studies in which spectroscopic, microwave, and other techniques are combined with shock tube and shock tunnel methods in efforts to unfold the behavior of the atoms and molecules of air.

CORNELL AERONAUTICAL LABORATORY, INC.
of Cornell University

J. P Rentschler
CORNELL AERONAUTICAL LABORATORY, INC.
Buffalo 21, New York

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City ..... Zone..... State.....

☐ Please include employment information.

MB

All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin.

NEWS
New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 120A)

The Model 166C Display Scanner provides an output to duplicate, on an X-Y recorder, any repetitive waveform appearing on the CRT trace. The permanent, large-scale record obtained provides higher resolution than either the CRT display or a photograph. The trace can be viewed on the CRT while an X-Y recording is being made.

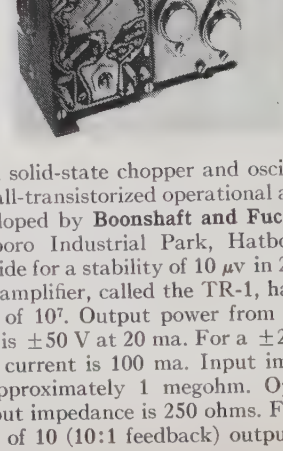
Model 166C converts high speed signals to slower signals having the same waveshape. The unit's scanning speed is arranged to keep the Y output within the bandwidth of conventional recorders.

The other new plug-in, Model 166D Sweep Delay Generator, delays the main sweep of the 160B and 170A oscilloscopes for detailed examination of a complex signal or pulse train. In addition, the instrument offers a unique mixed sweep feature to show an expanded waveform segment while still retaining a presentation of earlier portions of the waveform. Thus, both low speed and high speed phenomena can be observed on the same trace.

Model 166D's delay time is 1 microsecond to 10 seconds. Delaying sweep is in the 18 calibrated ranges from 2 microseconds/cm to 1 second/cm in 1, 2, 5 and 10 sequence. Delayed length is 0 to 10 cm.

The Model 166C Display Scanner is priced at $300. Manufacturer is currently quoting delivery 10 weeks after receipt of order. The Model 166D Sweep Delay Generator is priced at $325 and is currently available from stock.

High-Gain Operational Amplifier



A solid-state chopper and oscillator in this all-transistorized operational amplifier developed by **Boonshaft and Fuchs, Inc.**, Hatboro Industrial Park, Hatboro, Pa. provide for a stability of 10 μv in 24 hours. The amplifier, called the TR-1, has a loop gain of 10^7. Output power from 0 cps to 1 kc is ± 50 V at 20 ma. For a ± 25 V output, current is 100 ma. Input impedance is approximately 1 megohm. Open loop output impedance is 250 ohms. For a loop gain of 10 (10:1 feedback) output imped

(Continued on page 124A)

EXECUTIVE OPPORTUNITIES

***Mgrs. & Executives** to $28,000*

***Engrs. & Scientists** to $19,500*

Communications
Microwave Comp. & Syst.
Solid State Circuitry
Counter Measures
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needs electronic engineers

for Jansky & Bailey Division's expanding antenna systems design activities and Solid Propellant Division's instrumentation needs.

ANTENNA ENGINEERS

For J&B's Research and Engineering Department: one, to head up group for shipboard antenna problems; the other to perform theoretical and practical calculations for high frequency antenna systems design. Experience in model measurements, antenna impedance, broad band H.F. and V.H.F. antenna design, and matching network design, B.S., M.S. in engineering or physics.

ELECTROMAGNETIC THEORY

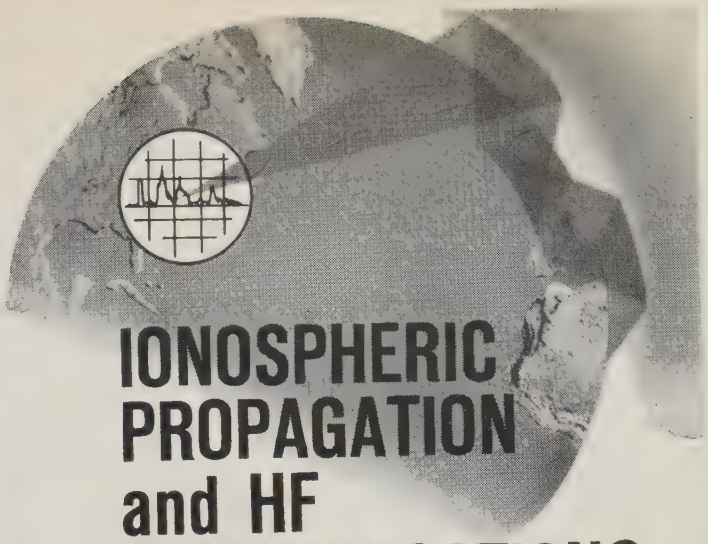
For J&B's R&E: to investigate radiation phenomena, theoretical calculations on antennae design and performance, wave propagation over frequency range LF to UHF and application of electromagnetic field theory. M.S. in EE or physics, 3-5 years experience.

INSTRUMENTATION ENGINEER

For ARC's rocket production & test facility: to maintain and develop instrumentation for firing bay measuring devices and propellant processing. Familiar with general measuring circuitry, pressure transducers, strain gages, multiple-point temperature recorders, electronic process instrumentation. B.S. engineering, 3 years practical experience.

Interested? Send your resume to Professional Personnel Recruitment (ire)

ATLANTIC RESEARCH CORPORATION
(Suburban Washington, D.C.)
An equal-opportunity employer



IONOSPHERIC PROPAGATION and HF COMMUNICATIONS

If you have the background, the imagination and the desire to contribute to important programs in these fields, you are invited to join a carefully selected team of outstanding scientists and engineers now contributing significantly to current knowledge through advanced research.

These programs are being conducted at our ELECTRO-PHYSICS LABORATORIES in the SUBURBAN WASHINGTON, D. C. area, ideally located from the viewpoint of advanced study which may be conducted at one of several nearby universities; for readily available housing in pleasant residential neighborhoods; and for the general amenities of living offered by this important Metropolitan center.

All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin.

OUR PRESENT NEEDS ARE FOR:

SENIOR IONOSPHERIC PHYSICISTS Ph.D. preferred, with several years' experience in the study of Ionospheric phenomena. Should be familiar with present knowledge of upper atmosphere physics and possess an understanding of current programs using rockets and satellites for studies in F-region and beyond. Qualified individuals with supervisory abilities will have an exceptional opportunity to assume project leadership duties on HF projects already under way involving F-layer propagation studies backed by a substantial experimental program.

SENIOR DEVELOPMENT PHYSICISTS Advanced degree in Physics or E.E. preferred. Must be familiar with latest techniques in the design of advanced HF receivers and transmitters and possess working knowledge of modern HF networks employing ferrites and metallic tape cores. Strong theoretical background in modern linear circuit theory desired. Will carry out laboratory development and implementation of new HF communications systems.

SENIOR ELECTRONIC ENGINEERS Advanced degree in E.E. preferred. Must be familiar with conventional pulse circuit designs and applications. Technical background should include substantial experience in data process and data recovery systems using both analog and digital techniques. Knowledge of principles and application of modern information theory including correlation techniques helpful. Will be responsible for the design of sub-systems.

JUNIOR ELECTRONIC ENGINEERS To assist Senior Engineers and Scientists in the development of HF communications and data process equipment. Should have formal electronics schooling and 2 years' experience in circuit design checkout or analysis of HF communications, Radar Pulse, Analog/Digital or Data Recovery equipment. Construction of prototypes of new and interesting equipment and design of individual components of communications and data processing systems will comprise the major efforts of selected applicants.

FIELD STATION ENGINEERS B.S.E.E. or equivalent, consisting of combined civilian or military technical school, with work experience. Presently employed as a field engineer or project engineer with a valid 1st or 2nd Class FCC license and a good command of some of the following: Radar, preferably high power; HF long-distance communications systems; Tropospheric or Ionospheric scatter systems. Must be willing to accept assignments in areas where dependents are not permitted for periods of up to one year. Differential paid for overseas assignments.

For a prompt reply to your inquiry, please forward resume in confidence to:
W. T. WHELAN, Director of Research & Development

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ACF INDUSTRIES

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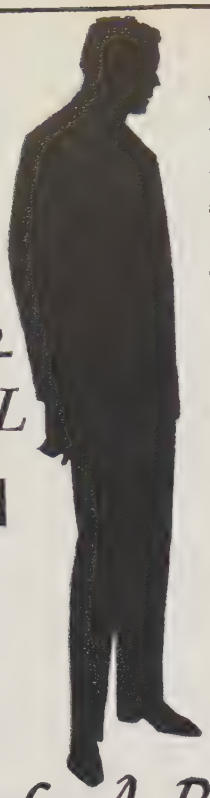

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NEWS New Products

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(Continued from page 122A)

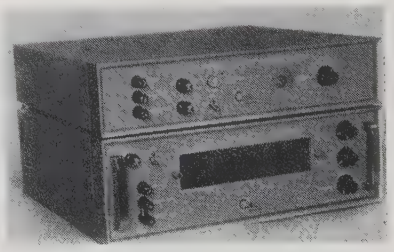
ance is 0.1 ohm. An automatic short circuit protection circuit is used. The complete unit, with case, weighs less than two pounds. Application is for measuring circuits, servo systems, analog computers, etc.

Thermal Reference Junction

Astra Technical Instruments Corp., 12930 Panama St., Los Angeles 66, Calif., announced recently the production of a new, standard line thermocouple reference junction, the RJ-24. A temperature control device, it is small in size ($3\frac{1}{4}\times3\frac{1}{4}\times5\frac{1}{4}$) and weighs less than four pounds. It provides a constant thermocouple reference temperature of 250°F with a 24 channel capacity. Uniformity between junction is 1°F. Ripple is less than ±1°F. Regulation throughout the environmental temperature range is less than 3°F, (−65°F to +165°F). Nominal power consumption is less than five watts. MIL-Spec type connectors of proper thermocouple materials are used. Other reference temperatures are available. Astra Tech also produces airborne units of 2, 4, 8 and 12 channel capabilities. Prices will be provided upon request. Address inquiries to R. L. Hixson, at the firm.

AC/DC Digital System

An ac power converter engineered for use with a V-45 dc digital voltmeter, suppling a precision ac/dc digital system, has been announced by **Cubic Corp.**, 5575 Kearney Villa Rd., San Diego 11, Calif.



The new converter, named the Model AC-45, is 19″ wide, $3\frac{1}{2}$″ high, and 14″ deep, and is designed to be twin mounted with its companion V-45 dc digital meter, also

(Continued on page 126A)

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More than 70 such programs are now on the boards at Northrop, with many challenging problems still to be solved, and new areas of activity constantly opening up for creative research. Whatever your specialty, drop us a line now if you want to find out more about the Northrop challenge.

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Fundamental investigation into related areas enabled e.m.r to undertake development of the test instrumentation subsystem for the Dyna-Soar manned space glider. This and other missile-satellite programs have created challenging opportunities for engineers with broad capabilities who seek an environment for maximum responsibility in probing beyond the range of present knowledge.

ELECTRONIC SYSTEM DESIGN ENGINEERS

Must have experience, state-of-the-art knowledge and a keen interest in the whole systems field, from initial concept through detail design and prototype development. Includes such areas as magnetics and solid state circuitry, RF transmitters and semiconductor applications, DC through UHF.

Please direct inquiries in confidence to ROLAND E. HOOD, JR., Personnel Manager, Systems Division

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NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 124A)

developed by Cubic. The combination of the ac/dc components provides a complete high accuracy measuring system for approximately $1500.

The Model AC-45 converter has a floating input which may be utilized up to 500 vdc above ground. Common mode rejection of 70 db at 60 cps is provided between signal ground and chassis ground. Rejection decreases roughly 6 db per octave at higher frequencies.

The AC-45, when used with the companion V-45 digital voltmeter, will permit measurement of voltages from 1 millivolt to 1099.9 volts ac in three manual ranges, to an accuracy of ±0.25% of full scale over a frequency range of 60 cps to 5 kc.

The manual ranging converter is simple and foolproof in operation. Input circuitry is protected against over-voltage and the simple control covers the complete range between 1 millivolt and 1099.9 volts ac.

A power receptacle is provided at the rear of the case for operating the companion digital voltmeter. On each range the instrument provides a full scale dc output of 10 volts and may be utilized to drive a recording pen or equivalent device having a floating input and suitable filtering.

The AC-45 unit weighs 17 pounds, is transistorized and uses one vacuum tube to provide maximum stability and low noise.

Directional Couplers

Sanders Associates, Inc., 95 Canal St., Nashua, N. H., has introduced a line of precision made TRI-PLATE® variable directional couplers for strip transmission line circuits with coupling as tight as 2 db's. The devices, which have broadband applications, can perform as adjustable directional couplers, power dividers or variable attenuators with controllable power handling capacity and wide attenuation range.

The new series of nine models of TRI-PLATE variable couplers are available from stock in three overlapping frequency ranges, each covering one or more octaves

(Continued on page 128A)

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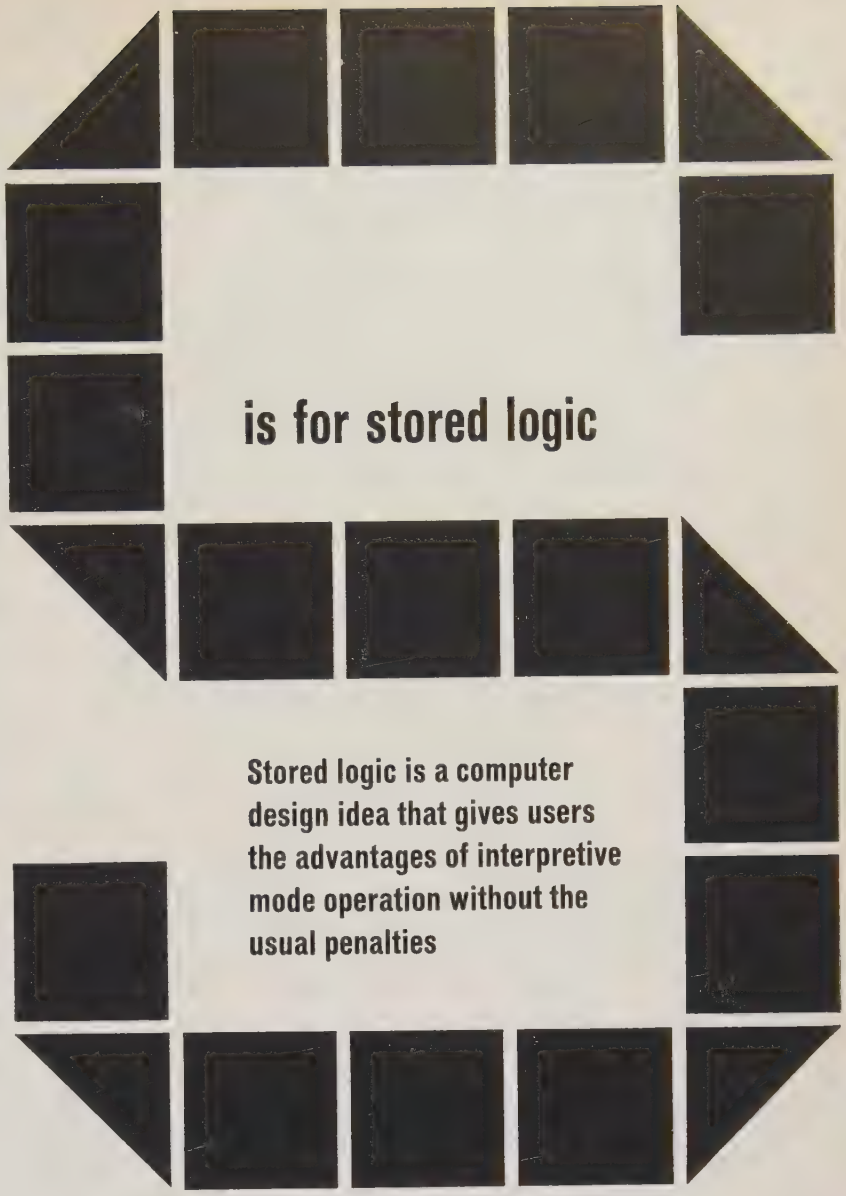
is for stored logic
Stored logic is a computer design idea that gives users the advantages of interpretive mode operation without the usual penalties

Computer people will admit that when a general purpose machine is operating in an interpretive mode it may spend most of its time doing non-productive work loosely called "housekeeping" functions. There's nothing wrong with this, as long as a user is willing to sacrifice efficiency for convenience. However, there are users and users, and many of them resent having to accept penalties of any kind—for any reason. Others need to have a computer that will perform as efficiently in many different modes as it does in the characteristic mode for which it was designed. Ordinary interpretive systems simply can't do this.

■ The basic trouble stems from the fact that computers and people (not counting programmers) don't speak the same language. The user often has to "command" his machine in codes archly called pseudo-instructions. The computer can't work from pseudo-instructions, but has to translate them into logic steps that will open gates, start counters, trigger flip-flops, and otherwise cause the machine to react to the problem.

■ It is this roundabout way of doing things that gives interpretive systems their bad reputation. (Small wonder, when a routine may take an ordinary computer ten to fifteen times as long to run interpretively as it would if the computer were lucky enough to be working from instructions originally wired into its back panel!)

■ It would be nice if all instructions could be wired into the computer in exactly the form that every present and future user would possibly desire. Of course, that's just dreaming. There are always new users and new problems.

■ A practical approach is to wire in only the basic (or generic) elements of all instructions, in order to get machine efficiency—and store in the computer's memory the logic for putting together the endless varieties of instructions and routines that users inevitably think up once they get their hands on a new computer.

■ This is the approach used by Ramo-Wooldridge in the design of its "stored logic" computer. The first model in R-W's stored logic line is the AN/UYK-1, an inexpensive machine being built for the Navy for general shipboard use. To illustrate its efficiency: the AN/UYK-1 can do a 15-bit basic add in 12 microseconds; a 30-bit arithmetic sine instruction, composed from some 30 generic commands, will take about six milliseconds. With 54 types of basic command elements, most of which have 32 options, the AN/UYK-1 offers its users more than 1100 generic portions (called logands, from LOGic commANDS) from which to compose instructions. These tailor-made instructions can be used directly by the programmer to solve his problems.

■ Recent contract awards for stored logic computers have created opportunities for adventurous programmers, systems analysts and circuit design engineers interested in helping to exploit the inherent advantages of the Stored Logic concept.

■ All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin.

Contact Mr. R. L. Eddy at

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NEWS New Products

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(Continued from page 126A)

between UHF and upper S-band. Within each frequency band three models are available with coupling at 3, 8, or 20 db's. TRI-PLATE variable directional couplers are also ideal laboratory tools and are furnished to mate with the complete line of over 300 TRI-PLATE modules or with transitions for use with coax.

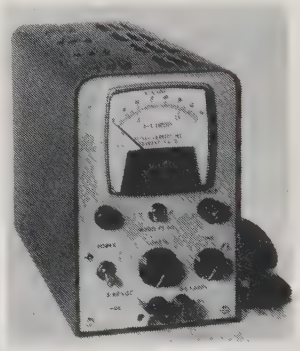
As a single unit the TRI-PLATE variable directional couplers provide equal 3 db outputs, or any other designated power ratio. Multiple units can be cascaded to provide adjustable power dividers with 4, 8, 16, or more adjustable (non-lossy) outputs in any ratio or power.

When used as a two port device the directional couplers function as variable attenuators with alterable power rating (up to 100 watts). Two to 50 db attenuation range is possible over the very broadband frequency range up to 3 octaves and more.

For complete information on the directional couplers contact product manager at the firm.

Solid State Power Supply

A line of new, solid state, regulated power supplies is being manufactured by **Flite-Tronics, Inc.,** 3312 Burton Ave., Burbank, Calif., in three main models which yield respective outputs of 3 to 20 vdc at 0–1.5 amperes; 5 to 60 vdc at 0–1.5 amperes; and 10 to 110 vdc at 0–.75 amperes. All models are provided with overload protection.



With an input ac power requirement of 105 to 125 volts at 60 cps, the power supplies measure $5\frac{1}{2}''$ in width, $8\frac{5}{8}''$ in height and 13″ in depth.

Among their features are: an output impedance of 500 milliohms or less dc to 2,000 cps; regulation for input line voltage variation of 15 millivolts or $\frac{1}{2}\%$, whichever is greater; ripple of less than 3 millivolts RMS or 0.05% RMS, whichever is greater. So as not to cause a hot spot chassis, unit may be operated in an ambient temperature of 85°C.

Resolution is refined to the point where the output dc voltage with the use of the fine control, may be set to any absolute voltage within 10 millivolts.

(Continued on page 130A)

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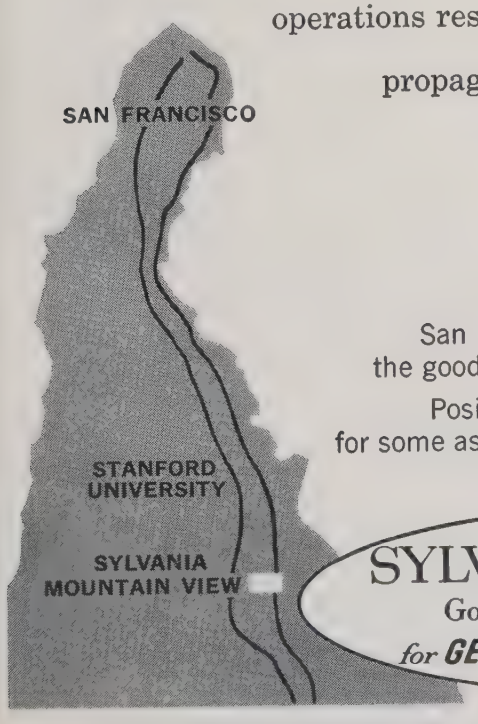
operations research ● systems analysis and synthesis

propagation studies ● signal analysis ● antennas

receivers ● transmitters ● data handling

product engineering ● mechanical design



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NEWS New Products

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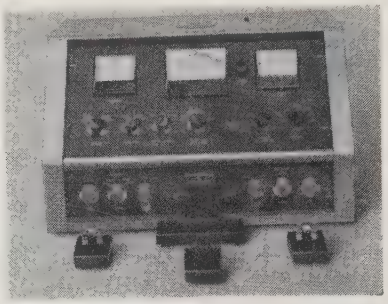
(Continued from page 128A)

Semiconductor Test Set

A general purpose semi-conductor test set for measuring the important ac and dc parameters of transistors, diodes, and rectifiers has been introduced by **Molecular Electronics, Inc.,** a subsdiary of **Precision Circuits, Inc.,** 87 Weyman Ave., New Rochelle, N. Y.

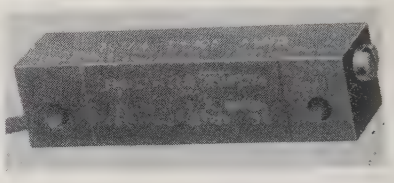


This new laboratory test and evaluation instrument can be used with all semiconductor devices, power as well as low level, silicon or germanium. Called the Model G-320, it offers PNP and NPN polarity, is simply calibrated and easily operated, permits rapid device insertion, and employs long-life semi-conductorized circuitry.

The Model G-320 measures dc current gain, h_{FE} 0–200 at the various base current levels of 25 μa, 250 μa, 2.5 ma, and 25 ma. Magnitude of ac current gain is measured at 1,000 cps. Price is $2,145. Full specifications and descriptive literature are available from the firm.

Carbon Potentiometer

The Trimpot® Division of Bourns, Inc., 6135 Magnolia Ave., Riverside, Calif., announces the availability of a new high temperature, humidity-proof carbon potentiometer, the Model 3011. This new, all plastic Trimpot potentiometer offers outstanding humidity resistance without the use of encapsulation. It can be used in potted module applications, even under stringent conditions of reduced vacuum and low viscosity potting materials without danger of leakage through the potentiometer enclosure. The Resiston® carbon element, a high temperature carbon on an inert and moisture-proof ceramic base provides good temperature and moisture stability.



(Continued on page 133A)

MANAGEMENT

Continued expansion by the Automatic Equipment Division has created two outstanding openings. This division is located in Northwest Chicago and is a major new industrial machinery division. These positions offer an extreme challenge to those who are resourceful and who can implement their creative ideas into action.

MANAGER-ENGINEERING

Graduate Electrical or Mechanical Engineer, preferably with graduate or multiple degrees, to be responsible for all engineering activity including development design and application engineering for a newly organized product department. Must have at least 10 years management experience with overall project level responsibility for the design and development of highly engineered systems. Should have a strong background in several of the following areas: logic circuitry, servo networks, hydraulic servo controls, tape recorders. Must also be experienced in the design of practical hardware for such systems, and should have demonstrated leadership and managerial capabilities.

CHIEF SYSTEMS ENGINEERING

Graduate Electrical Engineer to assume responsibility for highly engineered systems comprising logic circuitry, servo networks, hydraulic servo actuators, tape recorders and similar sub-systems. Will work in liaison capacity with development engineers, manufacturing, customers and field service groups. MS in electrical engineering desired but most important is demonstrated ability to analyze and trouble-shoot hydraulic servo controlled-electro mechanical systems.

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Personnel Manager

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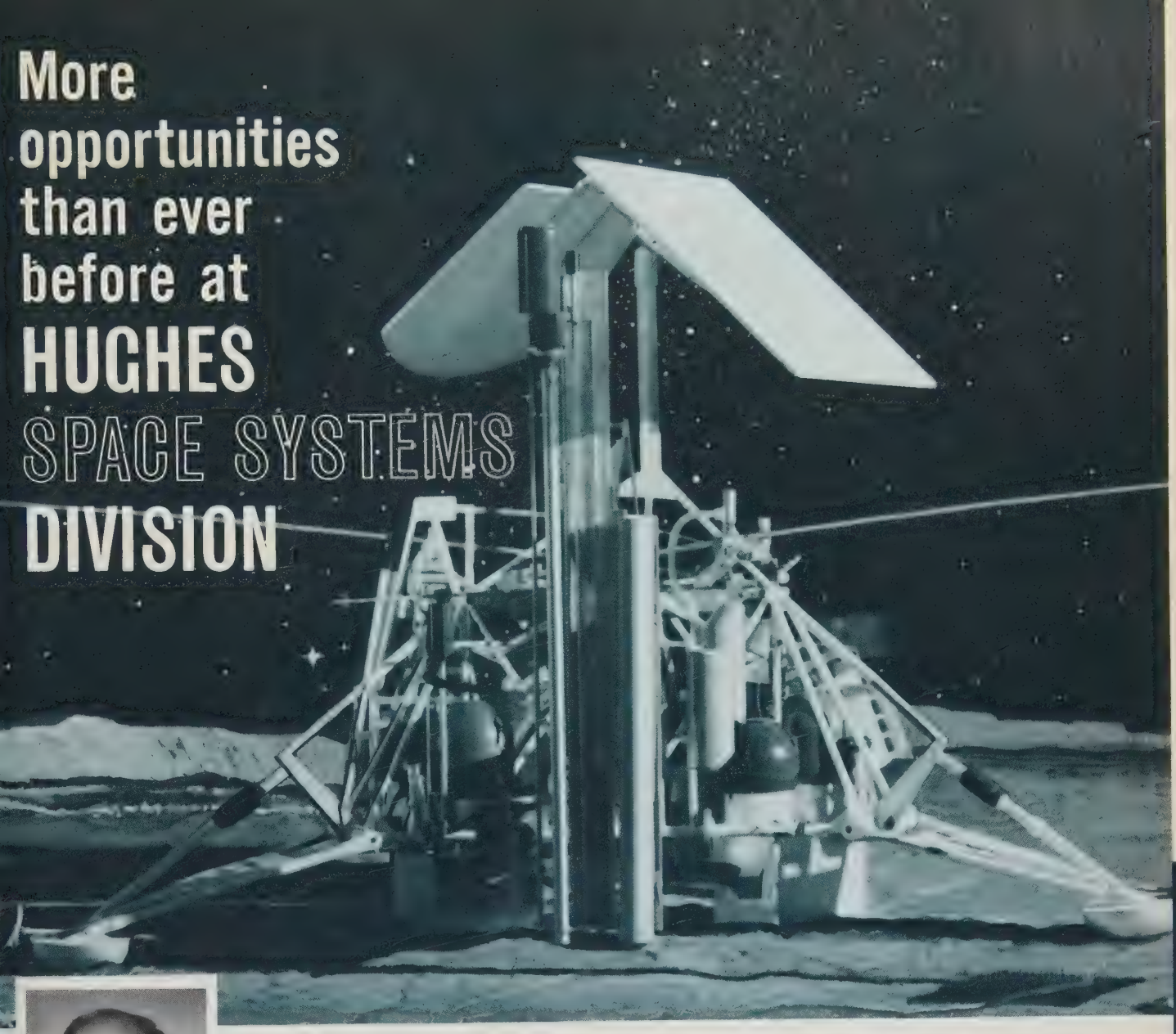

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The success of your professional career depends on your abilities, your training *and* the proper opportunities to apply these. Such opportunities now exist through the unusually challenging and responsible positions available in the Space Systems Division of the Hughes Aircraft Company.

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Openings include development of electronic propulsion systems for space probes. Positions call for capability of working in either a project or research capacity on direct energy conversion and storage systems. Openings exist for both junior and senior engineers. An advanced degree in Electronics or Physics is preferred.

Armament Control Analysts

Experience in the synthesis of integrated fire control loops including the tie-in of sensing and guidance equipment is desirable. A strong background in target data sensing devices, particularly at optical wavelengths, is very helpful, as is experience with computers.

Instrumentation Engineers

Involves the integration of advanced instruments into spacecraft such as Surveyor. Includes design for in-flight reliability; technical direction of instrument subcontractors; development of test programs and test equipment; and determining instrument interactions. Experience with subcontract procedures is also helpful.

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NEWS New Products

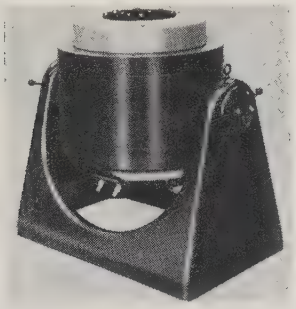
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 130A)

Specifications are resistances, 20K to 1 Meg; resolution, infinite; power rating, 0.25 watt at 50°C; operating temperature range, −65 to +150°C; humidity, meets MIL-STD-202A, Method 106; mechanical life, 500 cycles without discontinuity; weight, approximately 0.1 oz.; size, 1″ ×0.28″×$\frac{5}{16}$″; and price, $5.20 to $4.33 in typical usage quantity.

Electrodynamic Shaker

Ling Electronics Div., Ling-Temco Electronics, Inc., 1515 So. Manchester, Anaheim, Calif., introduces a new electrodynamic shaker, Model 300, with 5,000 pound force rating and built-in "piggy-back" capability for environmental testing.



Basic shaker needs no accessories for piggy-back application and features Ling's closed-loop cooling system for field and armature structures.

Armature weighs 49.5 pounds, with coolant, providing for larger payloads. Maximum load for 10 g vector is 450 pounds, for 20 g vector 200 pounds. Maximum acceleration is 100 g's. Frequency range is 5 to 3,000 cps. Low voltage requirements of armature make Model 300 ideal for evacuated chamber applications.

In "piggy-back" chamber, hermetically sealed shaker operates in following environmental ranges: Temperature, −100 to +350° Fahrenheit; humidity and altitude, unlimited. Same ranges apply to full chamber installations, with full Thermal Barrier Accessory.

Loop type flexures with stiffness of 1,000 pounds per inch offer maximum lateral restraint, linear spring constants, and 1 inch peak-to-peak continuous duty stroke.

Price and availability data may be obtained at factory sales office.

Microwave Signal Generator

A new "S" Band (3 kmc) stabilized signal generator providing full-octave bandwidth is now available from **Strand Labs, Inc.,** 294 Centre St., Newton 58, Mass.

(Continued on page 134A)

ENGINEERS

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Send us 3 complete resumes, telling us your present and desired salary; the kind of work you want and where you would like to live. That is all you have to do!

THEN YOU—

Wait to hear from us or our clients. There is no need to write directly to any companies, as we do all that for you and at absolutely NO COST TO YOU!

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- ☐ Receivers
- ☐ Analog and Digital Devices
- ☐ Radar Techniques
- ☐ Logic Design
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Current Openings with

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SINGLE SIDEBAND COMMUNICATIONS

Requires 6 to 10 years experience, with emphasis on high power vacuum tube amplifier design. Also desirable is a background in single sideband linear amplifier techniques.

LOGIC DESIGN

To participate in system design of all-electronic communication switching systems. Will be responsible for solving circuit problems from concept and interpretation of laboratory test results to final approved drawings. Requires BSEE with 5 to 10 years experience in telecommunications or other switching field.

RELIABILITY

1/ To set up a reliability group in military electronics. Requires BSEE with 10 years experience in electronics engineering, including design, and a minimum of 3 years in reliability.

2/ To make reliability analyses of circuits and components for commercial electronic telecommunications. Requires BSEE with 5 years experience in engineering design and 3 years in reliability.

ANTI-SUBMARINE WARFARE

To supervise receiver section of ASW Laboratory. Requires BSEE with over 5 years experience including RF and audio circuit design, ASW communications problems and signal processing.

ELECTRONIC SWITCHING

For development work in electronic telephone switching and other areas of digital logic, storage and control. Requires BS in EE or Physics with several years experience and a sound knowledge of solid state.

Please address resumes to M. J. Downey.

GIIIIIID

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All qualified applicants will be considered for employment without regard to race, creed, color or national origin.

ELECTRONICS ENGINEERS *WITH* RADAR EXPERIENCE

Silver Springs, Maryland

Armour Research Foundation, a leading independent research organization, offers exceptional long term opportunities for experienced electronics engineers who wish to work in Silver Springs, Maryland (convenient to both Washington and Baltimore areas). The Foundation is now forming a highly competent technical group to aid in the development of an advanced radar-missile system. Present assignments involve primarily circuit design and experimental system work. Applicants must have at least a B.S. degree in Electrical Engineering with an above average academic record and several years of radar experience.

Work on this program offers exceptional opportunities for professional and personal growth. Staff members receive attractive salaries, up to four weeks vacation, generous insurance and retirement benefits, and tuition paid graduate study. All qualified applicants will receive consideration for employment without regard to race, creed, color, or national origin. Please reply in confidence to Mr. S. M. Ream.

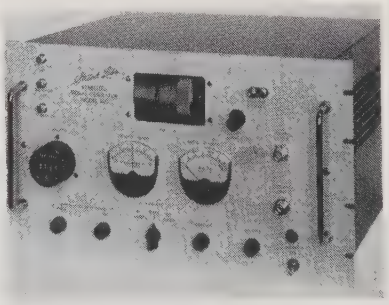
ARMOUR RESEARCH FOUNDATION

ELECTROMAGNETIC COMPATIBILITY ANALYSIS CENTER
N.E.E.S. ANNAPOLIS, MARYLAND

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 133A)



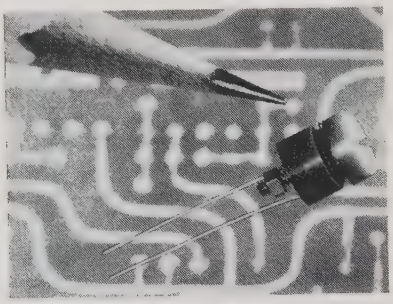
Designated as Model 800, this new instrument is suited for use as a prime signal with varactors and other frequency multiplying systems, as well as for test and calibration wherever high RF stability is required.

The instrument's design incorporates a patented microwave discriminator and high-gain dc feedback amplifier providing automatic frequency control of the klystron oscillator to 0.01 parts per million, and frequency setability accuracy of 0.05%.

The Model 800 stabilized signal generator is available in both a standard desk cabinet and a relay rack mounting unit.

Potentiometers

Miniature Electronic Components Corp., Holbrook, Mass., manufacturer of potentiometers, announces two new additions to its line of microminiature trimmer controls. Models PC-1 and PC-2 are wirewound units available in resistance values from 100 ohms to 10K. Both the stud-mounted PC-1 shown in the illustration and the lead-mounted PC-2 have a diameter of $\frac{1}{4}$ inch and a body length of $\frac{1}{4}$ inch.



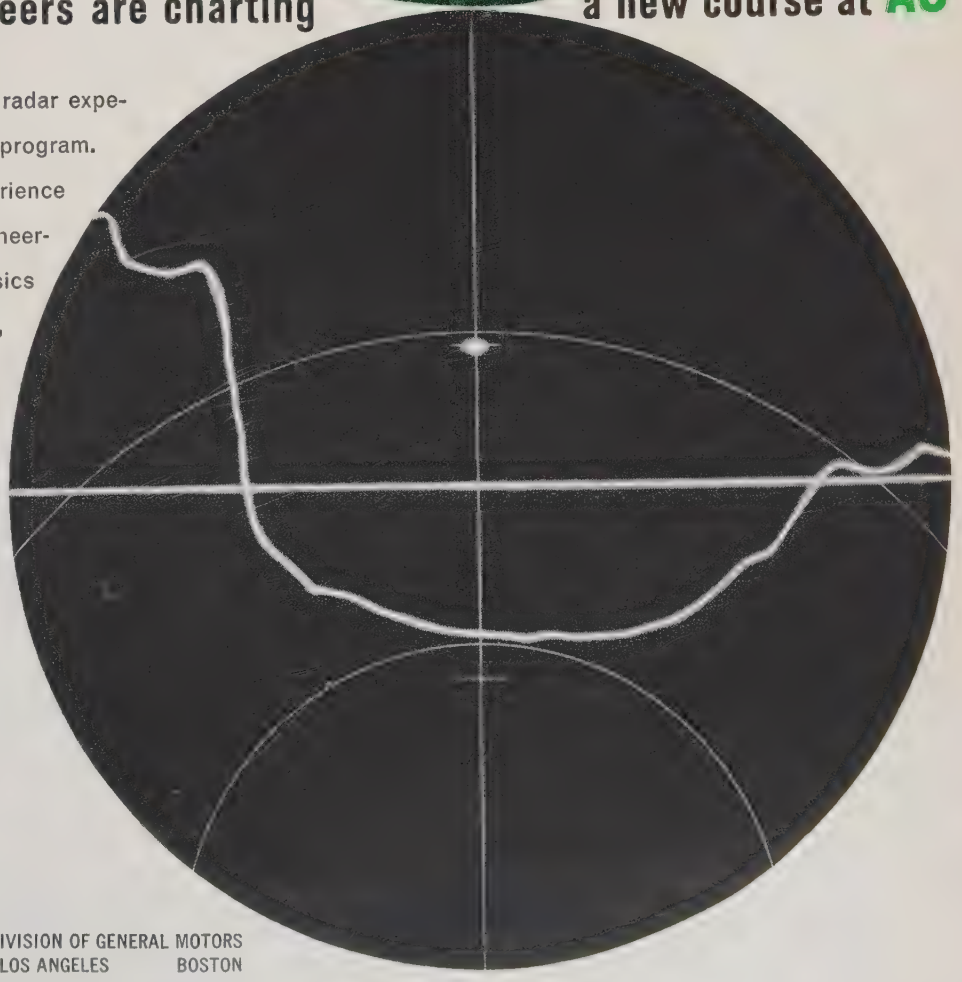
Specifically designed for potting, these low-cost units employ open construction such that the potting compound can penetrate the potentiometer to fix the wiper in any pre-adjusted position and render the unit immune to shock, vibration, and humidity.

Specifications include an end resistance of less than 1%, a resolution of 0.25% to 1% depending upon resistance value, a power rating of 0.25 watt, and a unit weight of 0.02 ounce.

(Continued on page 136A)

AC engineers are presently developing an improved Bombing Navigational System (BNS) that will enable the B-52C&D to fly low-level, high-speed bombing missions—regardless of terrain. The Air Force has assigned AC the responsibility for Systems Integration of the B-52C&D BNS. This responsibility will include program and engineering integration, and coordination of the associate contractors involved in the production phase. . . . . .

radar systems engineers are charting a new course at AC



AC is seeking qualified men with radar experience to work on this important program. If you have radar systems experience and a BS or MS in Electrical Engineering, Mechanical Engineering or Physics please contact Mr. G. F. Raasch, Director of Scientific and Professional Employment, Dept. 5753, 7929 S. Howell, Milwaukee 1, Wisc.

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AChiever Inertial Guidance Systems for Titan II, Thor and Mace.
Bombing Navigation Systems for B-52C&D and B-47.
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- HUMAN ENGINEERING OF CONTROLS AND DISPLAYS
- DIGITAL COMPUTER DEVELOPMENT
- MICROWAVE EQUIPMENT DEVELOPMENT
- ELECTRO-OPTICAL DEVICE DEVELOPMENT

All positions require a minimum of six years' experience in military research and development activity and an M.S. or Ph.D. degree in Electrical Engineering or Physics. All qualified applicants will be considered regardless of race, creed, color, or national origin. Replies to this ad should be addressed to: T. F. Wade, Technical Placement, The National Cash Register Company, Main and K Streets, Dayton 9, Ohio

I want a "Whatcha-ma-call-it?"
It's a "thinga-ma-jig that . . . . . "

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THE INSTITUTE OF RADIO ENGINEERS
1 East 79th Street, New York 21, N.Y.

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 134A)

Datapulse Transformers

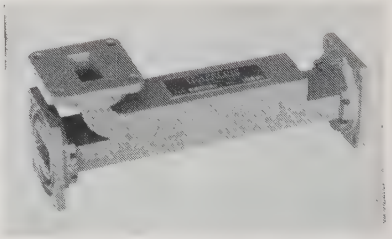
Claimed by the manufacturer to bring "a new dimension to the science of transformers," the new series of Datapulse transformers introduced by **PCA Electronics, Inc.,** 16799 Schoenborn St., Sepulveda, Calif., is designed to specifications which enable engineers to predict in advance the behavior of these transformers to a degree of accuracy never before achieved.

Datapulse Transformers are unique in that they utilize a "standard" equivalent circuit to reduce parameter variations to manageable proportions. These proportions may be plotted on a series of graphs and charts supplied by PCA. By the application of known circuit requirements to these charts, the engineer may then determine the exact pulse transformer best suited to meet those requirements.

A general information catalog on the parameters covered by Datapulse Transformers may be obtained from the firm.

Directional Couplers

Waveline, Inc., Caldwell, N. J., has designed a group of miniature 6 db waveguide directional couplers which exhibit good coupling characteristics over a broad band frequency range.



These couplers combine small physical size with a minimum of coupling variation over a large portion of the waveguide frequency range.

To illustrate the degree of performance attained, we have listed below X-band and Ku-band performance data. It should be noted that the coupling variation over the indicated frequency range did not exceed ± 0.5 db in either version of miniature coupler.

(Continued on page 138A)

Philco WDL has immediate and challenging engineering positions open in Discoverer, Midas and Advent programs in these categories:

Data systems engineering

Technical Staff — communications and data systems

Systems test planning

Systems reliability analysis

Design Engineering — surface and vehicle electronics

Human factors and operations analysis

Field tracking station activation

Tracking and control systems design

Logistics and station support

And others

All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin; U. S. citizenship or current transferable Department of Defense clearance required.

if your future is for growing...

People at Philco's Western Development Laboratories are proud of their achievements, secure in the steady growth of their company, enthusiastic about their future.

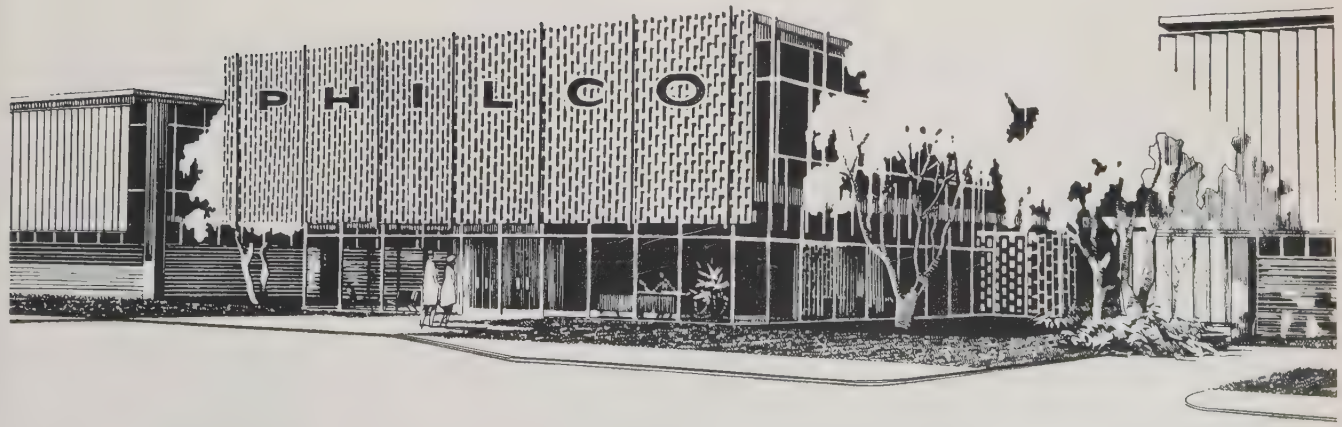
In just four years, Philco WDL has expanded from a staff of 18 to more than 2,000. At the end of this year WDL will open a new 250,000-square-foot facility which will accommodate a continually expanding complement of engineers, scientists and supporting staff members.

Philco WDL, the space organization which designed and built the Courier satellite, conducts an ever-growing development program in tracking, satellite instrumentation, communications, data processing and command—moving hand in hand with the federal government in space exploration and space age defense.

Your growth with a growing company, ideal living on the Northern California Peninsula, professional and monetary advancement commensurate with your own ability—these are some of the advantages and satisfactions of working at Philco Western Development Laboratories.

If yours is one of the fields listed at the left, write today to Mr. W. E. Daly, in confidence, of course, Department R-10.

This new 250,000-square-foot addition to Philco Western Development Labs will open by the end of the year.



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Increased technical responsibilities in the field of range measurements have required the creation of new positions at the Lincoln Laboratory. We invite inquiries from senior members of the scientific community interested in participating with us in solving problems of the greatest urgency in the defense of the nation.

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• *A more complete description of the Laboratory's work will be sent to you upon request.*

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NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 136A)

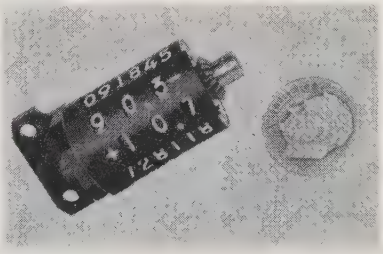
Additional special designs may be generated for miniature couplers with coupling values from 5 to 10 db. These special designs will cover a 12% bandwidth of frequency range with a maximum coupling variation of ±0.5 db.

Dimensions for the narrow or broad wall X-band coupler, 6 db are: length, 5¼ inches; frequency range, 8.5 to 9.6 kmc; coupling variation, ±0.4 db; directivity, 15 db minimum; vswr (any arm), 1.20 maximum

For the narrow or broad wall Ku-band coupler, 6 db they are: length, 5 inches; frequency range, 15.5 to 17.5 kmc; coupling variations, ±0.5 db; directivity; 12 db minimum; vswr (any arm), 1.30 maximum.

Add-Subtract Counter

The type 30–130 variation counter developed by **Melland Gear & Instrument Co., Inc.,** 88–06 Van Wyck Expressway, Jamaica 18, N. Y., is used for digital readout applications where plus and minus, right and left, or directional readouts are required. The counter, as shown, can be driven to ±99.9 and is prevented from overrunning by built-in mechanical stops.



This miniature instrument counter was developed for applications requiring a precision, high speed, low torque counter. Utilizing internal gearing and constructed entirely of precision machined parts, the counter can be mounted on a bracket as illustrated and driven by means of a through shaft from either side.

The diameter of the drums is 0.672 inches and the numerals are 0.125 high. The maximum torque is 0.25 oz-in. at room temperature. Running speed is 350 rpm continuous with intermittent operation at 600 rpm.

WWV Receiver

General Mills, Inc., 1620 Central Ave., Minneapolis 13, Minn., has a new transistorized WWV receiver, Model 925, which can be used in any application where the time standards, frequency standards or radio propagation notices of radio stations WWV and WWVH are required. Because of complete transistorization, high sensitivity and compact construction, the receiver is suited to field operations where a

ASTROPOWER

offers responsible positions to

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Ph.D. or M.S., with two or more years of experience, to participate in an advanced design and development program utilizing transistor and silicon-controlled rectifiers.

SOLID STATE PHYSICISTS

Ph.D. or equivalent, with experience in: preparative techniques and measurements of physical properties and a working knowledge of band theory and its application to solid state devices; or in design of solid state devices.

Astropower is an expanding propulsion R & D firm located in the ideally-situated research center at Newport Beach, California, one of the finest recreation areas in the world. Superior employee benefits. Submit resumes to Dr. George Moe, Vice President, Research. Qualified applicants will receive consideration without regard to race, creed or national origin.

ASTROPOWER, INC.
A subsidiary of Douglas Aircraft Co.
2968 Randolph Avenue
Costa Mesa, California

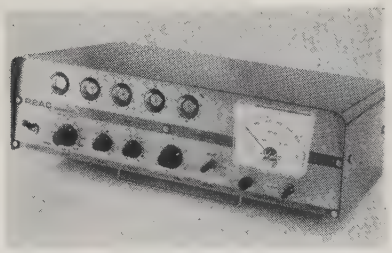
time standard is needed or where instruments and devices must be synchronized.

The receiver tunes by manually switching to the several transmitted frequencies, and contains provisions for separating the 600 and 440 cps audio tones. In addition, 1000 cps tics are processed so they can be used with a minimum of background interference. A separate RF input and a signal level meter facilitate zero-beat comparisons between the WWV carrier frequencies and a locally generated signal.

A descriptive eight-page booklet, containing detailed information and specifications, can be obtained by writing to the advertising manager at the firm.

"Analytical" Scaler

A new, transistorized "analytical" scaler, compatible with all types of counting systems, from basic GM counting to automatic sampling systems, has been announced by **Radiation Equipment & Accessories Corp.**, 665 Merrick Rd., Lynbrook, N. Y.



The Model E-130 represents application of the latest solid-state techniques to the field of nuclear science. Its high voltage supply allows wide range in addition to providing a fixed reference source voltage for accurate reproducibility, a plus-factor in scintillation counting.

Separate gain and discriminator controls facilitate precise integral spectrometry. When coupled with the firm's transistorized preamplifier, millivolt sensitivity is obtained for proportional gas flow counting.

The E-130 is available at $795 FOB, Lynbrook. Complete specification sheet available on request.

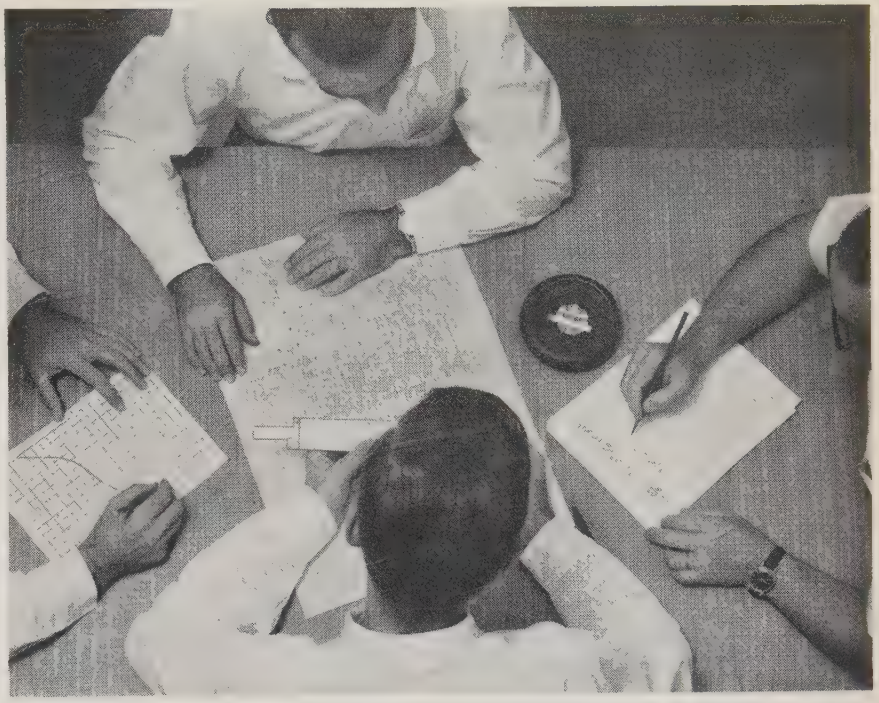
Pulse Simulator

Telemetrics, Inc., 12927 S. Budlong Ave., Gardena, Calif., recently announced development of its Series 500 digital electronic signal simulator as a laboratory tool for use in digital and computer circuit design. Flexibility permits simulation of digital communications systems, memory exerciser and as a programming device for computers in data reduction centers.

(Continued on page 140A)

at MOTOROLA in PHOENIX:

Leadership in action



Project leaders at Motorola are engineers and scientists who have won their right to leadership — with ideas, talent, and dedication — while working on challenging assignments in Project Groups. These teams are carefully selected — to get the optimum combination of experience and capabilities, and to utilize each man's talents to the fullest. Opportunities for individual advancement and career fulfillment are broadened and enhanced by Motorola's widely-diversified activities in advanced military electronics. Leadership in action is a vital force at Motorola. It provides the stimulus for challenging the state-of-the-art in Project Group assignments, assuring engineers and scientists of ever-broadening horizons for tomorrow. And Arizona's world-famed climate and beauty add to the whole family's year-round health and pleasure.

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- Digital Circuitry Design
- Microwave and Radar
- Reliability and Components
- Solid State Devices

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Write Phil Nienstedt (Dept. 1001)

MOTOROLA

MILITARY ELECTRONICS DIVISION, WESTERN CENTER

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Chicago, Illinois, and at Culver City and Riverside, California

ELECTRONIC ENGINEERS

Research and Development

Armour Research Foundation, one of the largest independent research organizations, offers exceptional opportunities in Chicago for qualified Electronics research engineers to work on a variety of industrial and military research projects in the fields of

RADAR

ANTENNAS

CIRCUIT DEVELOPMENT

MICROWAVES

GUIDANCE & CONTROL

PROPAGATION

COMMUNICATION

SIGNAL PROCESSING

SEMICONDUCTOR DEVICES

MEASUREMENT TECHNIQUES

Minimum requirements include a B.S. degree in Electrical Engineering with above average academic grades. Research experience in one of the above fields is highly desirable, though not absolutely necessary.

Our environment combines the better features of academic and industrial research and offers you the opportunity to work on advanced research programs.

Staff members receive attractive salaries, up to 4 weeks vacation, generous insurance and retirement benefits, and tuition paid graduate study. Our Chicago location offers excellent cultural and recreational advantages. All qualified applicants will receive consideration for employment without regard to race, creed, color, or national origin. Please reply in confidence to Mr. S. M. Ream.

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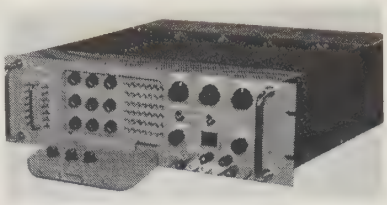
OF ILLINOIS INSTITUTE OF TECHNOLOGY
TECHNOLOGY CENTER, CHICAGO 16, ILL.

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 139A)



The all solid state ESS-500 presents data in a serial form of any type of digital coding simulating digital output of a multiplexer or digitizer in RZ, NRZ and modified NRZ formats for checkout of telemetry systems.

Optional features include parallel outputs, increased word and frame lengths, sub-commutation, and bit rates up to a megacycle.

For additional information, contact Robert L. Burr at the firm.

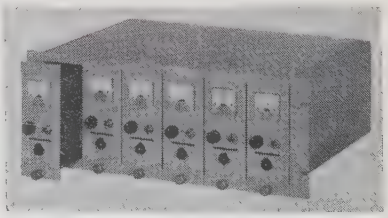
Wickham Joins Ampex

Dr. Donald Wickham has been appointed manager, Material Research and Development for **Ampex Computer Products Co., Components Div.,** P.O. Box 329, Culver City, Calif.

In 1950 he received his MS degree in chemistry from the University of Denver where he had also completed his undergraduate work. He continued his education at MIT receiving his PhD in 1954. Dr. Wickham remained with MIT another six years working in the Chemistry Section of the Computer Components Group, Lincoln Laboratory, the last three years as leader of the Chemistry Section.

Dr. Wickham is the author of several articles on the properties of magnetic oxides. He is a member of the ACA (American Crystallographic Association).

DC Amplifier



Model 94G, differential DC amplifier developed by **Video Instruments Co., Inc.,** 3002 Pennsylvania Ave., Santa Monica, Calif., serves a dual purpose. It is an isolation amplifier with unity gain—and it is a

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ASW SYSTEMS
RADAR BEACONS & DISPLAYS
FIELD ENGINEERING

Our modern facilities are located in residential suburbs of Washington, D. C. and New York City, affording you a choice of city, suburban or country living. Houses and apartments in all price ranges are readily available. Advanced study under tuition refund may be conducted at several nearby universities.

All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin.

For details, contact:
A.C. Sugalski
Mgr., Professional Staffing, Dept. A-1
(at our Riverdale facility)

ACF ELECTRONICS
DIVISION

ACF INDUSTRIES

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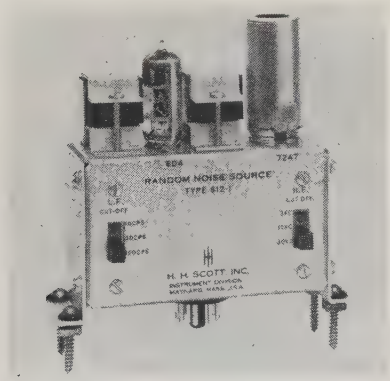
high current amplifier capable of driving low sensitivity, fluid damped galvanometers. Through usage of a guard shield, the 94G rejects common mode signals by a ratio of 100,000,000:1. Since the input is fully isolated from the output, the 94G eliminates the frequently encountered "ground loop" problems: For driving galvanometers, the 94G provides up to 200 ma of output current, thus acting as an effective impedance match between high impedance voltage sources and low impedance galvanometers. The bandwidth of this amplifier is 10 kc. Chopper stabilization minimizes long term drift. The design allows operation with long cable lines, both at the input and output.

Random Noise Source

H. H. Scott, Inc., Instruments Div., 111 Powdermill Rd., Maynard, Mass., manufacturer of sound measuring and analyzing instruments, announces a new series of random noise sources which provide random "white" noise with Gaussian amplitude distribution where a limited degree of flexibility is required. The units utilize an octal socket for both standard console power input and random noise output.



By use of two 3 position frequency cutoff switches, nine selected frequency ranges of noise are available. The 812-1 is designed primarily for vibration work, while the 812-2 is useful for general audio frequency investigations. The 812-1 and 812-2 are priced at $95 each and may be delivered from stock. Special models of the 812 incorporating any three selected lower and upper cutoff frequencies from 3c to 50 kc are available.

The 813 series provides output in any single fixed frequency range desired (within 3c to 50 kc) with a cutoff rate of 6 or 12 db per octave. Output will vary from 1v RMS to 3v RMS depending on the frequency range selected.

The 814 series are engineered and produced to order. They provide a noise output with shaped spectrum for simulation of specific noise environments. The 814 will find application in telephone line testing, telemetering, probability generation, acoustic analgesia, production component testing, creating speech interference and masking signals, and simulating ambient sea noise.

For complete technical specifications and pricing information write to the firm.

(Continued on page 142A)

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. . . you'll find it at Electronics Park

At General Electric's headquarters for electronics in Syracuse, New York, you'll discover projects in every phase of modern electronics, aimed at every kind of application — Consumer, Industrial, Military.

These projects are distributed over research laboratories, development and design groups, manufacturing facilities and field service centers. Different operations are structured to provide independent, localized management yet all offer technical cooperation to each other. In the next building — or the next office — you'll be able to pinpoint a technical expert for consultation on any electronic problem.

THE CHART BELOW offers a broad indication of areas in which we are actively engaged. Check off those areas that are of interest to you — fill in the coupon — and mail to the address below. We will advise you of specific opportunities now available at General Electric's Electronics Park.

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COMMUNICATIONS						
VLF Systems	•	•	•	•		
Secure Communications	•	•	•	•		
Underwater Acoustic Communications	•	•	•	•		
Telecommunications	•	•	•	•	•	
DATA RECORDING AND DISPLAY						
Equipments		•	•	•	•	•
Thin Film Cryogenics	•		•			
3-D Display	•		•			
DEFENSE SYSTEMS (active, passive)	•	•	•	•		
Missile Guidance and Navigation	•	•	•	•	•	•
Air Weapons Control	•	•	•	•	•	•
Aerospace	•	•	•	•		
COMPUTERS (applications)	•	•	•	•	•	•
MATERIALS AND PROCESSES (Electrochemical, Organic, etc.)	•		•		•	
SEMICONDUCTORS —circuitry, devices, equipment	•		•	•		
UNDERSEA ACOUSTIC SYSTEMS						
Transducer & Techniques Development (L. F.)	•	•	•	•	•	
ASW and Surveillance Development	•	•	•	•	•	
TELEVISION (Broadcast & Industrial)						
Receivers	•	•	•	•	•	
Transmitters (AM, FM, and TV)			•	•	•	•
Studio and CCTV Equipment			•	•	•	
TUBES						
Entertainment	•		•	•		
Indicator, Display and Pickup Devices	•	•	•	•		
Infrared and Electron Optics	•	•	•	•		

To: GENERAL ELECTRIC CO. • Electronics Park, Syracuse, N.Y.
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Degree: _____ Date: _____

U. S. citizenship or a transferable Department of Defense security clearance is required in the military electronics areas. All communications will be held in strict confidence.

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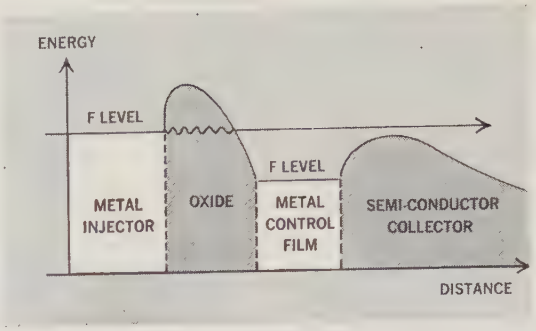
ELECTRONIC AMPLIFICATION BASED ON

HOT ELECTRONS in METAL FILMS

Philco scientists recently have verified that energetic electrons injected into thin metal films are capable of being separated from normal conduction electrons clustered about the Fermi level. Potential barriers at the boundaries separating the film and insulating layers prevent conduction electrons from escaping while hot electrons possess sufficient energy to surmount the barriers. Low impedance injection and high impedance collection of hot electrons is the basis of a new class of solid state devices —the Metal Interface Amplifier. Injection is accomplished by quantum mechanical tunnelling through a thin oxidic layer on one face and collection by an adjacent germanium substrate. High collection efficiency implying low loss of electron energy through collisions is ensured by the use of films whose thicknesses are significantly less than the electronic mean free path.

This program is typical of research activities at Philco's new Research Center. For further information, please write:

Mr. W. P. Diefenbach
Personnel Director
Philco Research Center
Blue Bell, Pennsylvania



Electron potential diagram of thin-film structure.

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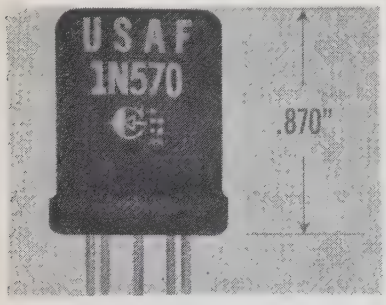
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 141A)

1500 Volt Silicon Rectifier

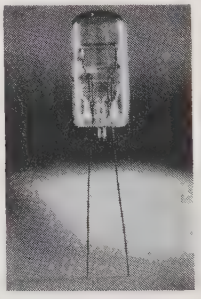
Columbus Electronics Corp., 1000 Saw Mill River Rd., Yonkers, N. Y., announces the availability of silicon rectifier type 1N570 USAF per MIL-E-1/1275. Designed to replace the 6X4 tube where ultra-reliability and miniaturization is required, the plug-in 1N570 is rated at 1500 piv and 37.5 ma per half wave section.



Columbus offers a complete line of KV silicon rectifier combinations including types for convection cooling, immersion in oil-filled containers and cast in epoxy blocks. They are part of more than 350 types made available by the Yonkers firm. For further information, write Gustav Bard, sales manager at the firm.

Ultra-Stable Crystal

An ultra-stable 5 mc, 5th overtone "Glasline" crystal, the JKG6AS, is announced by the **James Knights Co.,** Sandwich, Ill.



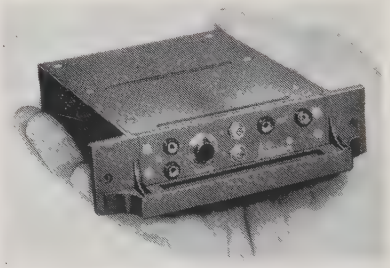
With an aging characteristic of 1×10^{-9}/day, this small, shock resistant crystal unit is designed principally for use in frequency standards exposed to rigorous environmental conditions.

Under vibration of 10 to 200 cps at 10G frequency change is less than 1×10^{-8}. The unit will withstand a shock of 100G when tested per MIL-STD-202A, method 202A.

The JKG6AS is avilable in a frequency range from 4.966 megacycles through 6.133 megacycles. Detailed specifications are available from the firm.

Pulse Rate Integrator

A new, compact, miniaturized pulse rate integrator specifically designed for telemetry of missile and airborne frequencies is announced by **Anadex Instruments, Inc.,** 14734 Arminta St., Van Nuys, Calif. Called the Model PI-300-J77 pulse rate integrator, it is used in missiles and aircraft for converting frequency to voltage for telemetering.



This latest Anadex pulse rate integrator is highly accurate and features a self-contained calibrating circuit as well a high-input and low-output impedances. It has a full scale output of 5 volts and is adjustable over a 4 to 1 frequency range. Input sensitivity is adjustable from 10 mv. to 1 volt.

Other features of the Model PI-300-J77 pulse rate integrator include high density packaging and plug-in module. The complete unit weighs only 20 ounces including case. It also features printed circuit construction and is potted and ruggedized to withstand shock and vibration. It has an operating temperature range of −55°F to +165°F. Delivery is 30 days after receipt of order.

For complete, detailed specifications and prices on the Model PI-300-J77 pulse rate integrator, write direct to the manufacturer.

(Continued on page 144A)

Demanding Assignments for
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SYLVANIA RSL

Reconnaissance Systems Laboratory on the
SAN FRANCISCO PENINSULA

SENIOR SYSTEMS ENGINEERS

System management responsibility for new programs at Sylvania's Reconnaissance Systems Laboratory offers the broad spectrum of challenging problems involved in large reconnaissance systems. A minimum of 5 years experience is required in progressive assignments leading to and including systems analysis, determination of functional systems requirements and synthesis and integration of subsystems with over-all program requiring knowledge of the capabilities and limitations of techniques and components. Should have background in analysis and synthesis of HF and VHF communication systems employing a variety of modulation techniques and/or digital data handling and automatic control systems. Advanced degree preferred.

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The design groups in RSL have openings for experienced engineers with five or more years of design and development background in one or more of these areas: communications receivers, digital tuning, RF distribution, DF equipment, magnetic tape recorders, teletype equipment, dynamic displays, demodulators, data transmission, data converters and computers. Opportunities for both technical specialists and supervisors.

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Complete information may be obtained by writing, in confidence, to Mr. W. L. Pearson at:

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ASSOCIATE DIRECTOR OF APPLIED PHYSICS

Will be leader of the Applied Physics Section of Advanced Development Engineering Group. This Section will be responsible for Applied Research in areas relating to Military Electronics, such as antennas, microwave circuitry, propagation, with some emphasis on solid state physics. Requirements are a PhD in Physics or Electrical Engineering with capabilities in some aspect of Electromagnetic field, and experience in technical leadership.

ASSOCIATE DIRECTOR OF APPLIED MATHEMATICS

Will be leader of Mathematics Section of Advanced Development Engineering Group with dual responsibility for conducting research programs in Mathematics and providing mathematical services to the Electronics Division. Problem areas relate to conception and design of new Military Electronics systems. PhD in mathematics and experience in technical leadership are required.

Technical opportunities exist in the following areas:

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Solid State Device Application
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Ocean Survey Systems Engineering

Send resume to: Mr. J. F. Caldwell, Dept. 435

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NEWS New Products

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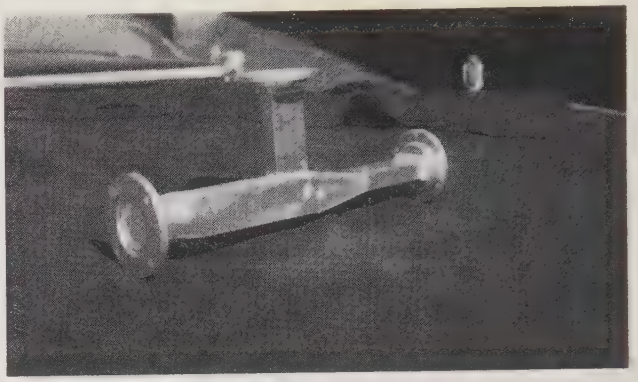
(Continued from page 142A)

Dual Polarized Adaptor for 6000 Mc Parabolas

The Model DPA-6000 Dual Polarized Adaptor, a product of **Mark Microwave, a Division of Mark Products Co.,** 5439 W. Fargo Ave., Skokie, Ill., is specifically designed for the Mark Microwave series of parabolic antennas such as Model P-6048, P-6072, P-6096, and P-60120. These represent 4-foot, 6-foot, 8-foot, and 10-foot diameter units for the 6000 Mc range. Units are available for the 5925 to 6425 Mc frequencies for common carrier applications and also for frequencies up to 8000 Mc for the other services in this range.



(Continued on page 146A)

TECHNICAL WRITER

Cornell Aeronautical Laboratory has an attractive opening for a qualified technical writer who would work closely with our technical staff in writing proposals and reports.

This position is at a professional level and provides a fine opportunity for the person chosen to contribute significantly to important research programs.

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EAST VS. WEST

DIAGRAMS VS. EQUATIONS

THE COMPUTER'S ANSWER TO A LONG-STANDING COMPUTER ISSUE.

For a decade East Coast and West Coast computer designers have been using different methods of representing computer logic—the Easterners with diagrams, the Westerners with equations.

$$LBSM = (LXA1)(LXA2*)(LFCA*) + (LXA1*)(LXA2)(LFCA*) + (LXA1*)(LXA2*)(LFCA) + (LXA1)(LXA2)(LFCA)$$

$$LFCAJ = (LXA1)(LXA2)$$

$$LFCAK = (LXA1*)(LXA2*)$$

In the example illustrated here, the diagram and the equation tell us exactly the same thing. Either represents a serial full adder where the sequence of pulses at the output, LBSM, will represent a serial binary number that is the sum of two serial binary input numbers occurring at LXA1 and LXA2. (The asterisks indicate binary complements; for example, whenever LXA1 is energized LXA1* is not, and vice versa. LFCA is a carry flip-flop.)

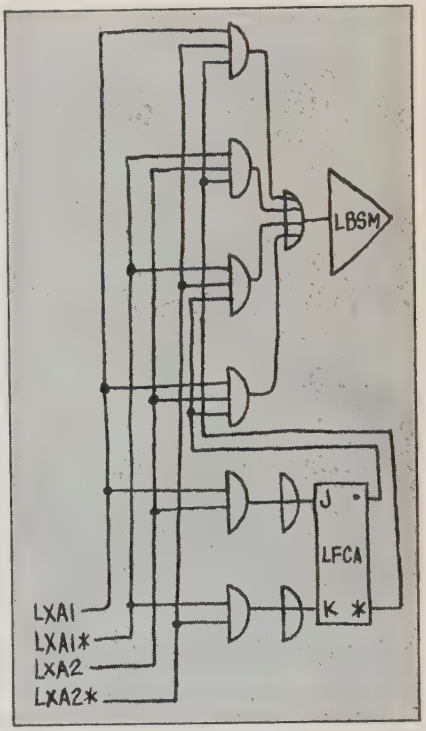
There are persuasive arguments on both sides. Eastern proponents of diagrams point out that the logical interconnections can be seen at a glance and followed through any number of stages by eye. The logical structure of an entire system can be understood from a diagram more directly and intuitively, they maintain, than from a set of equations.

The Western argument for equations goes like this. It's not true that diagrams communicate better to the viewer's intuition, except at first exposure. The human mind is highly adaptive. After working analytically with the equations for a while, the mind begins to operate intuitively in that symbology. Then the intrinsic superiority of equations over diagrams begins to make itself evident. One advantage, say the Westerners, is that equations can represent the same information more compactly and efficiently, as our illustration shows. Another is that equations lend themselves better to computer manipulation of logical design information.

As evidence of the latter advantage Westerners point to a recent achievement of some Litton Systems people: a completely mechanized procedure for translating logical designs into wiring lists, including operational simulation of the design to verify its accuracy. A procedure enormously facilitated by the computerizability of logical equations. It's easy to picture the benefits in cost, delivery schedules, reliability, price. Using only a partial development of this method Litton Systems recently brought a major computer system from concept to operation in less than a year.

Now under consideration at Litton: a machine that will accept as inputs a supply of standard computer components and a set of coded specifications defining the logical functions desired, and will crank out completely fabricated systems.

Maybe you think we've loaded the argument in favor of equations. You're right. But we're ready to listen to arguments on either side. Drop us a card. Or better still, drop in in person. You'll like the



imagination-stretching atmosphere generated by Litton management's appreciation of the rewards of creative controversy. And we have a few excellent opportunities for computer design people. Ask for S. L. Hirsch at Litton Systems, Inc., Data Systems Division, 6700 Eton Ave., Canoga Park, California.

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Also — Amperite Differential Relays: Used for automatic overload, under-voltage or under-current protection.
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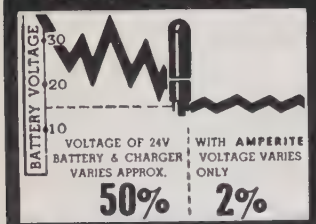
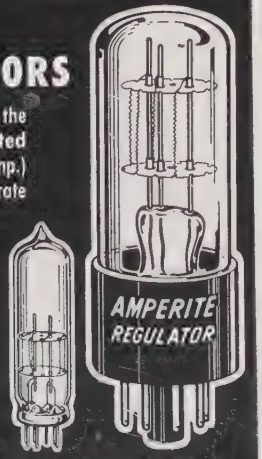
Amperite Regulators are designed to keep the current in a circuit automatically regulated at a definite value (for example, 0.5 amp.) . . . For currents of 60 ma. to 5 amps. Operate on A.C., D.C., or Pulsating Current.




Hermetically sealed, they are not affected by changes in altitude, ambient temperature (—50° to +70° C.), or humidity . . . Rugged, light, compact, most inexpensive . . . . . List Price, $3.00.
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NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 144A)

The series of parabolic antennas as listed utilize the (patent applied for) circular waveguide feed as described on page 7 of Bulletin 600 (available on request). The standard series such as the Model P-6048, 4-foot diameter parabolic antenna for the 5925 to 6425 Mc range comes equipped with circular waveguide feed and precision cast transition from circular-to-rectangular with standard UG-343A/U broad band choke flange input.

This circular-to-rectangular waveguide transition allows for full 360° adjustment of polarization without physical movement of the feed proper. This is accomplished thru an O-ring pressure seal and a swivel flange coupling. Thus, for single polarized applications, the standard parabolic antennas are normally used.

With the addition of the DPA-6000, Dual Polarized Adaptor, it is now possible to convert any existing Mark Microwave circular waveguide feed parabolic antenna in the field for dual polarized application. All that is necessary is to decouple and remove the precision cast circular-to-rectangular transition and replace it with the DPA-6000 Dual Polarized Adaptor. The DPA-6000 has two UG-343A/U broad band choke flange inputs for the two polarizations. Isolation between inputs is 35 db or better over the range.

While the two inputs launch waves into the circular waveguide which are orthogonal, the feature of continuous rotation is still maintained so that polarization adjustment is still available at the focus of the dish and this will allow for convenient field adjustment and peaking of both signals.

Thus, with the Model DPA-6000, for the first time it is possible to adapt an existing Mark Microwave field antenna installation from single polarized operation to diversity dual polarized system performance.

Facsimile System

New, high speed facsimile scanners and recorders capable of sending and reproducing graphic messages at 8 to 16 times present network speeds were demonstrated by **Alden Electronic & Impulse Recording Equipment Co., Inc.,** Alden Research Center, Westboro, Mass., at the 15th Armed Forces Communications and Electronics Association Convention in Washington, June 6–8.

The new units are designed specifically to take advantage of low-cost, broad channel communication links such as microwave and AT&T's "Telpak," now coming increasingly into use. Combined with presently-available tape storage and conversion equipment, the equipment makes possible whole new types of facsimile communication systems with reduced message costs.

The new units were described as "essentially the same as the present equipment operating on the Weather Bureau High Altitude Network at 120 rpm and other networks at 60 rpm, but with component changes which increase recording speed to 960 rpm.

"This means that using a 19 inch-wide recording head, standard for weather maps and said to be optimum for business use, the transmission time for two business letters or equivalent area is reduced from 8 or 16 minutes down to 2 letters per minute."

The Alden continuous flat copy scanner, which also was demonstrated, eliminates copy size restrictions by handling any width of copy in continuous lengths.

(Continued on page 150A)

Military–Commercial–Entertainment Rectifier Purchasers

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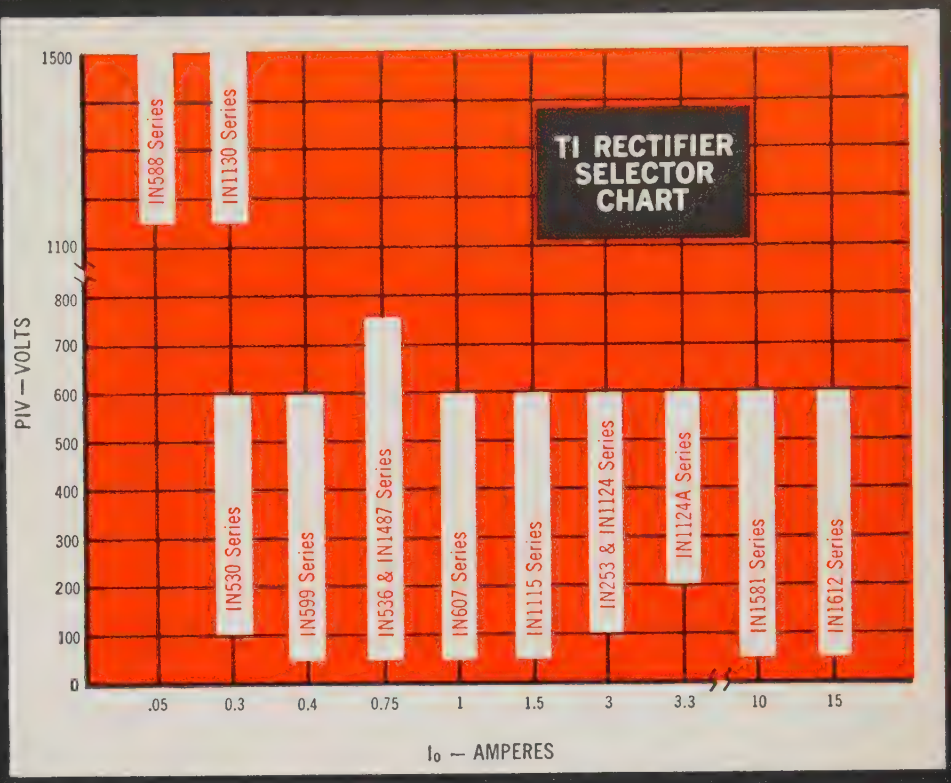

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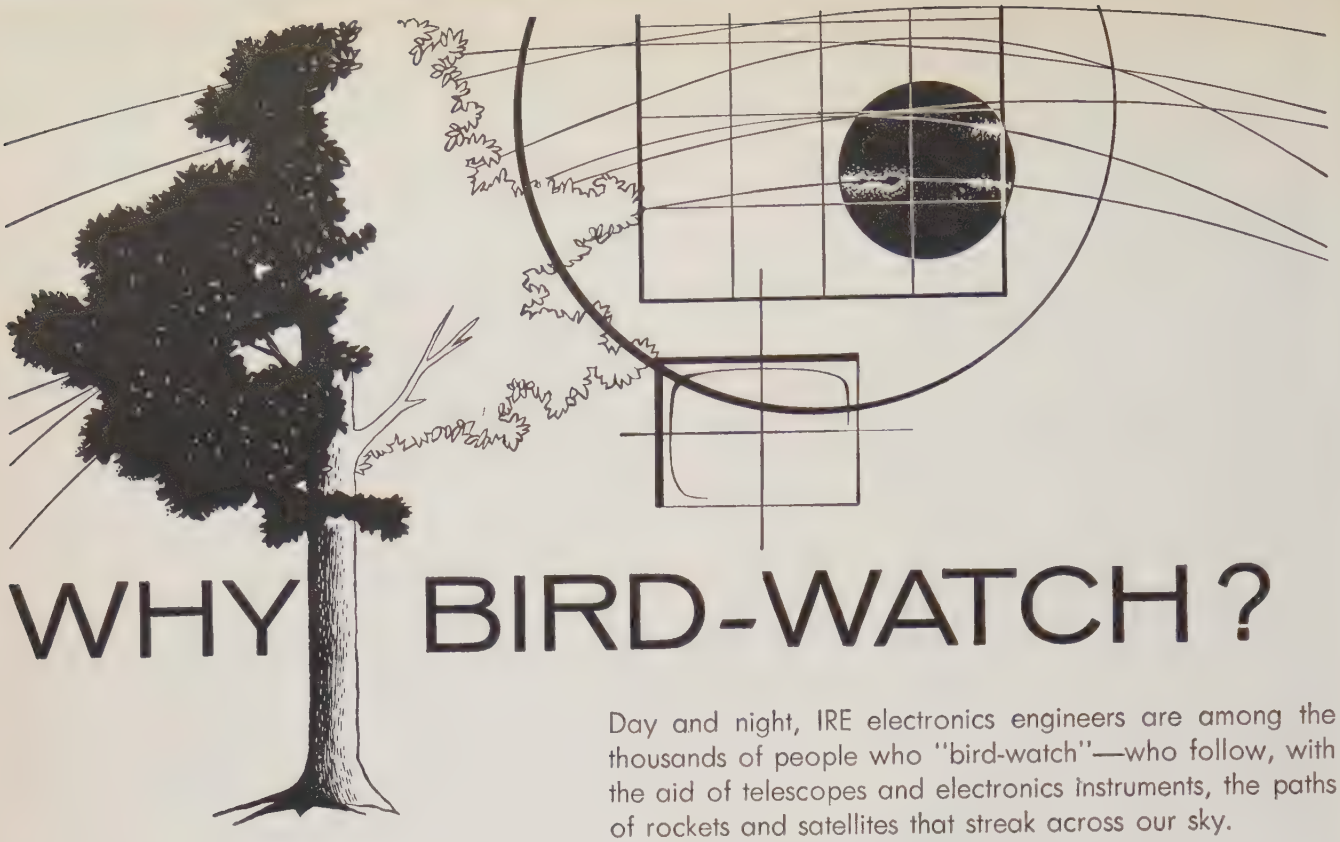

WHY BIRD-WATCH?

Day and night, IRE electronics engineers are among the thousands of people who "bird-watch"—who follow, with the aid of telescopes and electronics instruments, the paths of rockets and satellites that streak across our sky.

This bird-watching is vital to our nation's defense. But it is important, too, to keep an eye open at ground level—for tanks, and guns, and armored cars, and motorized infantry.

Television is being used more and more for military reconnaissance, and its effectiveness has been substantially increased by the researches of Peter C. Goldmark, of CBS Laboratories. At the same time, Mr. Goldmark has used his knowledge of TV to help us defend ourselves against other deadlier enemies—anti-bodies—by making possible the use of television in medical education.

Like other IRE members, Mr. Goldmark works continuously in a world of pure and abstract science; and again, like his colleagues, he relates his research to the world around us, to make a better life possible for all. The IRE salutes him with the Vladimir K. Zworykin Award for 1961, and commends to your attention the tremendous IRE membership of professional electronics engineers, whose zeal, intellectual ability and devotion to duty Mr. Goldmark so typifies.

The researches and findings of electronics engineers of Mr. Goldmark's calibre are published each month in *Proceedings of the IRE*. Your ad in *Proceedings* associates your company with a publishing project that benefits all mankind.

REMEMBERS THE MAN

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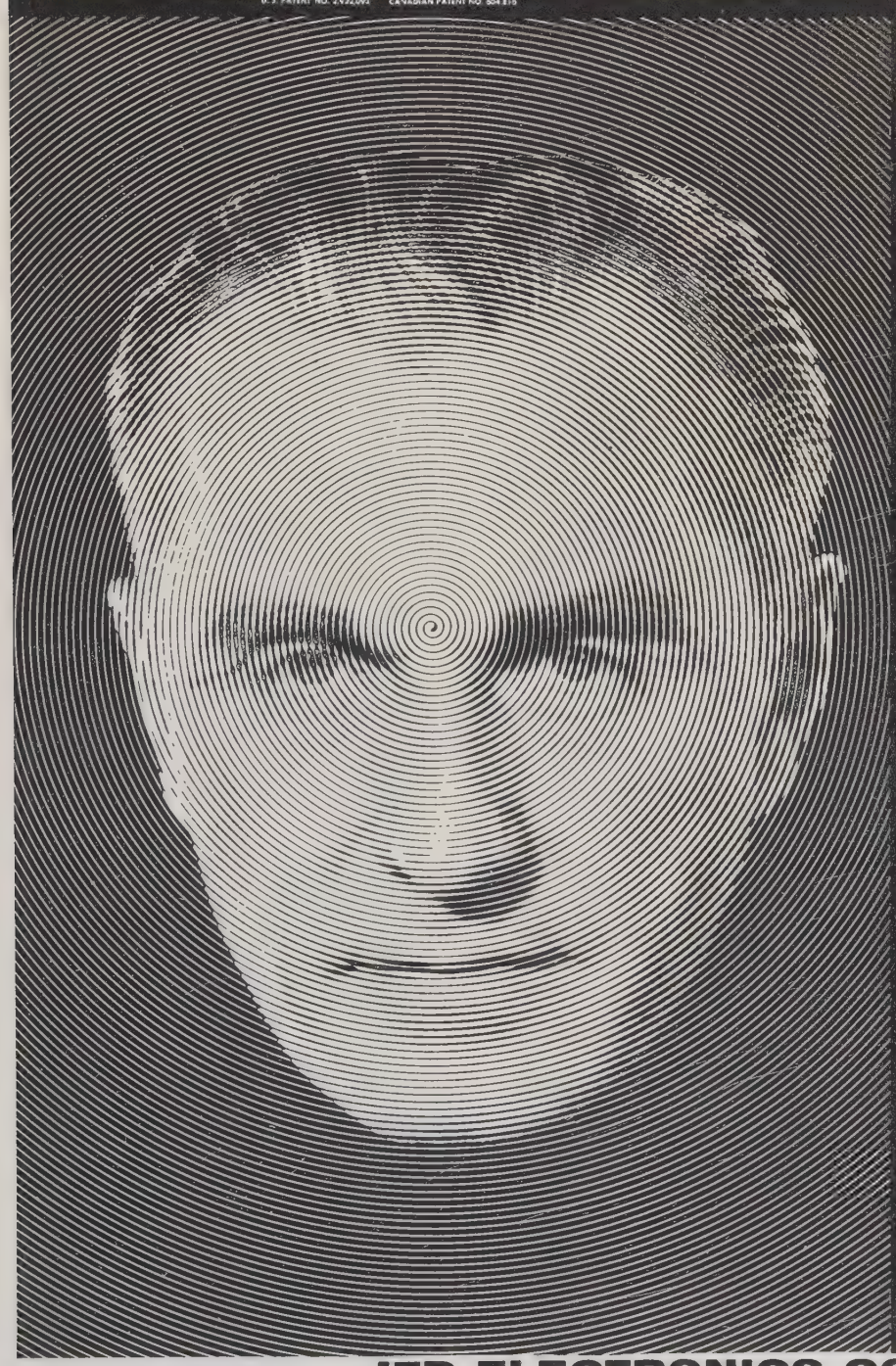

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(ACTUAL SIZE)

DIFFERENTIAL TRIMMER
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Features capacity that increases in one gang as capacity decreases in the other gang. The sum of the capacitance of both gangs remains constant in the peak-to-peak area of the tuning curve. Low expansion invar and precision bore glass provides increased stability across a wide operating temperature range (from −55°C. to +125°C.) 35 standard models available in panel mount and printed circuit types.

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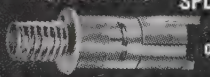

SPLIT STATOR TRIMMER
MODEL SP86G
Capacitance Range
0.8 to 4.2 pf plate to bushing
0.6 to 2.00 pf plate to plate

Features balanced electrode construction ideally suited for push-pull radio frequencies circuits, and similar networks where physical size is limited. When tuning, capacitance varies simultaneously from each plate to bushing, as well as plate to plate. 25 standard models are available in panel mount and printed circuit types.

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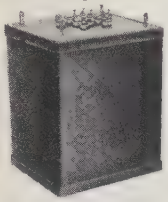
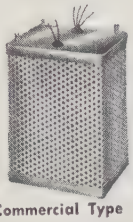

MIL Type


Commercial Type

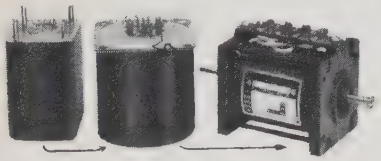
Here at last is a hermetically sealed magnetic voltage regulator that will provide constant output voltage regardless of line and/or load changes.

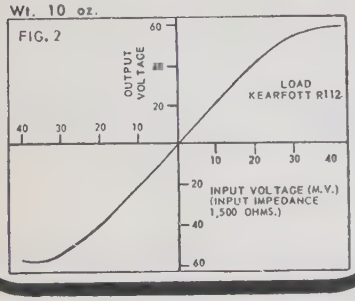
SUPPLIED EITHER MIL. OR COMMERCIAL				
CAT. #	INPUT VOLT.	LINE FREQ.	OUTPUT VOLT.	OUTPUT VA.
MCV- 620L	95-130 v	60 cps.	115	20
MCV- 670L	95-130 v	60 cps.	115	70
MCV-6130L	95-130 v	60 cps.	115	130
MCV- 670F	95-130 v	60 cps.	6.4	70
MCV-6130F	95-130 v	60 cps.	6.4	130
MCV- 420F	95-130 v	400 cps.	6.4	20

MAGNETIC AMPLIFIERS

- Hermetically Sealed To MIL Specifications
- No Tubes
- Direct Operation from Line Voltage
- Fast Response
- Long Life Trouble Free Operation
- Phase Reversible Output

Power Gain 2 x 10^8


Transistor Preamp. MAT-1 Wt. 10 oz.
Mag. Amp. MAF-5 Wt. 18 oz.
Motor



PARTIAL LISTING ONLY
WRITE FOR FURTHER INFORMATION ON THESE UNITS OR SPECIAL DESIGNS.
Send for NEW 48 page transformer catalog. Also ask for complete laboratory test instrument catalog.

FREED TRANSFORMER CO., INC.
1720 Weirfield St., Brooklyn (Ridgewood) 27, N. Y.

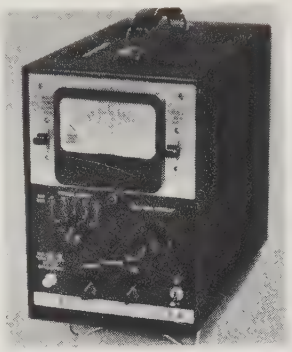
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 146A)

Microwave Power Meter

General Microwave Corp., 47 Gazza Blvd., Farmingdale, N. Y., announces the availability of the Type 451 microwave power meter. The instrument is designed to operate with a wide variety of 100 and 200 ohm, positive or negative temperature coefficient bolometers (e.g., barretters, thermistors and fuses).



Five scale ranges with full-scale readings of 0.1, 0.3, 1.0, 3.0, 10 milliwatts are provided. Corresponding decibel scales are −10, −5, 0, +5 and +10 dbm. Higher power levels may be measured using auxiliary calibrated attenuators or directional couplers.

Frequency range is dependent only on the exteral detector. Bias current to 16 milliamperes is provided.

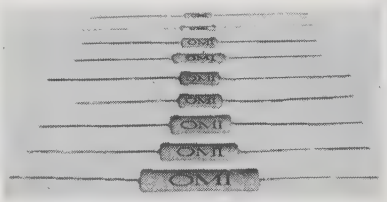
Accuracy is held to within ±5% of full scale reading.

Input power requirements are 115/230 volts ±10%, 50–1,000 cps.

The meter is available, in the portable cabinet-mount illustrated, from stock to 30 days at $235.00.

Resistor Line

Introduction of the series "T," TEMP-OHM resistor line has been announced by **Omtronics Manufacturing, Inc.**, P.O. Box 1419, Peony Park Station, Omaha 14, Neb.



Available in ten physical sizes, with wattage ranges from ½ to 10 watts rating, these components have a resistance range of from 0.05 ohms to 200K ohms and tolerances from ±0.05% to 5.0%.

Incorporating the use of TEMP-COTE high temperature coating materials allows

(Continued on page 152A)

MAJOR BREAKTHROUGH FOR HIGH and SUPERPOWER SSB

The Kahn Model SSB-58-1B high efficiency Class C SSB Exciter-Driver System, backed by more than 7½ years of continuous use, offers the first major breakthrough for high and superpower military and commercial SSB communications.

There are approximately seventy Kahn installations throughout the world, developing between 40 KW and 4 megawatts PEP.

Typical Kahn Model SSB-58-1A exciter-driver installation on 10 KW high-level AM Transmitter at Radio Swiss near Berne, Switzerland develops 40 KW PEP on SSB. (Picture courtesy of Radio Swiss and Brown, Boveri).

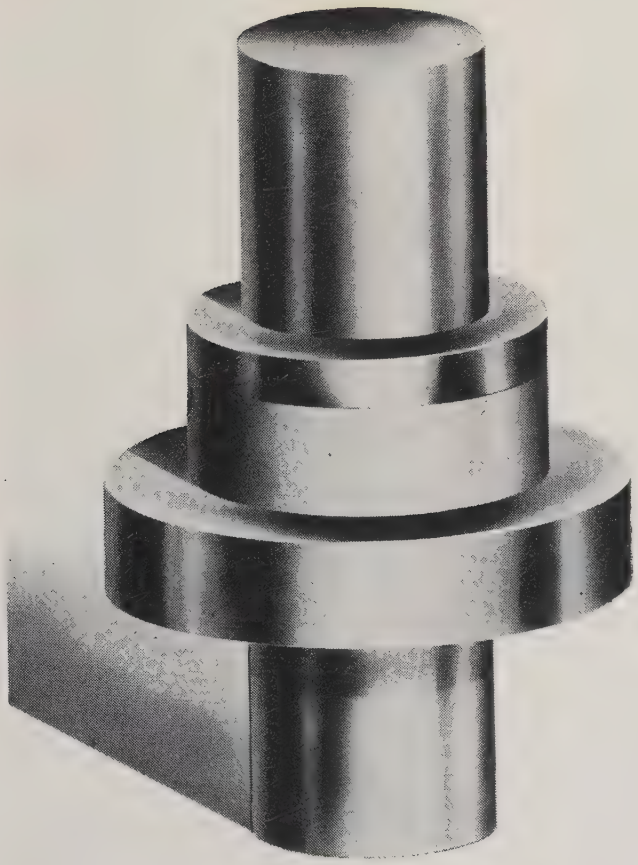
- 2½ times more power for a given plate dissipation — thus reducing overall SSB system cost, size, and weight.
- Greater sideband rejection at high and superpower levels.

KAHN RESEARCH LABORATORIES
KAHN RESEARCH LABORATORIES, Inc.
81 So. Bergen Place • Freeport, New York
FReeport 9-8800

IMMEDIATELY AVAILABLE

A COMPLETE LINE OF VARACTORS

in coaxial pin packages!



30 VOLT VARACTORS

These diffused junction mesa GaAs varactors are designed for use in harmonic generators and microwave switches. They feature high cutoff frequencies, low series resistance, low series inductance.

f_{co} @ — 6v	C_j 0.2 — 1.1 $\mu\mu f$ @ — 6v		C_j 3 — 6 $\mu\mu f$ @ — 6v		C_j 6 — 10 $\mu\mu f$ @ — 6v	
kmc min.		Rs ohms max.		Rs ohms max.		Rs ohms max.
10			MS2620	5.0	MS2630	2.5
20			MS2621	2.5	MS2631	1.3
40	MS2602	15	MS2622	1.3	MS2632	1.0
60	MS2603	12	MS2623	1.0		
80	MS2604	10				
100	MS2605	8				
120	MS2606	6				

Maximum reverse current 10 μa @ — 30v
Package capacitance typically 0.35 $\mu\mu f$
Series inductance typically 0.3 mμh
f_{co} calculated from Q measured at 10 kmc
Other capacitance ranges available on request

150 kmc VARACTORS

This series of diffused junction mesa GaAs varactors is designed for use in parametric amplifiers. It features high cutoff frequency, low series inductance, low reverse leakage, and may be cooled to liquid nitrogen temperatures.

Type	f_{co} @—2v min.	f_{co} @—6v min.	C_j Zero Bias min.	C_j Zero Bias max.
MS2501	60 kmc	80 kmc	0.2	1.1 $\mu\mu f$
MS2502	80	105	0.2	1.1
MS2503	100	133	0.2	1.1
MS2504	120	160	0.2	1.1
MS2505	150	200	0.2	1.1

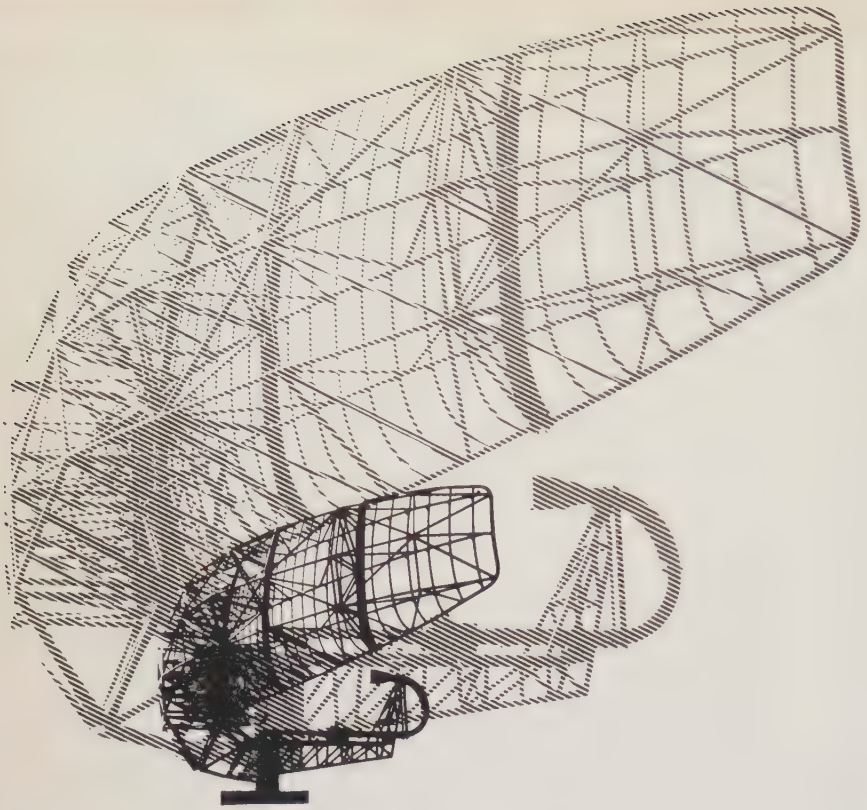
Maximum reverse current 0.10 μa @ — 6v
Package capacitance typically 0.35 $\mu\mu f$
Series inductance typically 0.3 mμh
f_{co} at minus 2 volts calculated from Q measured at 10 kmc
Other capacitance ranges available on request

FOR ADDITIONAL INFORMATION WRITE OR CALL

micro state electronics corporation

152 FLORAL AVENUE, MURRAY HILL, NEW JERSEY CR 7-6600

GALLIUM ARSENIDE MATERIAL • GERMANIUM TUNNEL DIODES • SOLID STATE UHF AND MICROWAVE AMPLIFIERS AND OSCILLATORS • GALLIUM ARSENIDE VARACTORS



SAFETY OF LIFE IN THE AIR

SELENIA AIR TRAFFIC CONTROL RADAR OFFERS:

SOLID, LONG RANGE, HIGH ALTITUDE DETECTION OF EVEN THE SMALLEST TYPES OF AIRCRAFT THROUGH THE USE OF HIGH AVERAGE POWER AND A PARAMETRIC AMPLIFIER RECEIVER.

THE MOST ADVANCED TRUE MTI SYSTEM IN THE WORLD WITH THREE PULSE CANCELLATION FOR CLUTTER-FREE PRESENTATION AND STAGGERED REPETITION RATE FOR BLIND SPEEDS ELIMINATION.

THE USE OF L-BAND ASSURES SHARP AZIMUTH DISCRIMINATION AND LOW PRECIPITATION ATTENUATION WITH REASONABLY SIZED ANTENNA STRUCTURES

SPECIAL HIGH-LOBE ANTENNA FOR GREATLY IMPROVED SUBCLUTTER VISIBILITY FOR CLOSE-IN TARGETS AND CIRCULAR POLARIZATION FOR NEARLY COMPLETE CANCELLATION OF RETURNS FROM PRECIPITATION.

A RADAR COMPLETELY COMPATIBLE WITH JET AGE REQUIREMENTS.

A DIRECT SUCCESSOR OF THE PROVEN RAYTHEON AIRPORT AND AIRWAYS RADARS CHOSEN BY THE AIR TRAFFIC CONTROL AUTHORITIES IN CANADA, THE UNITED STATES AND SWITZERLAND.

Your inquiry is invited for more specific information
ROME ITALY
P.O. BOX 7083

MANAGED BY RAYTHEON

SELENIA

Excellence in Electronics in Europe

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 150A)

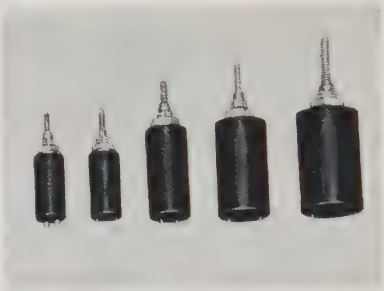
their use in temperatures to +350°C, and they are impervious to all effects of salt-spray, humidity or abrasion, in accordance with applicable paragraphs of MIL-R-26.

For printed circuit applications, series "TR" resistors, with radial leads, incorporate all the features found in the series "T" resistors, at no increase in price.

Availablity: Small quantities, one week or less; volume production, two to three weeks. Price: Ranges from $0.35 to $1.65, depending on quantity, resistance value and tolerance.

Tunable RF Coils

Relcoil Products Corp., Spring St. & Route 75, Windsor Locks, Conn., an affiliate of Hi-G, Inc., has introduced a new series of tunable RF coils, to be called the RC and RCP series.



The RC series and RCP (plug-in) provide a wide range of precision RF tunable coils in the medium and high Q areas. The inductance values range from 0.1 microhenry to 15,000 microhenries, in overlapping standard tuning ranges.

Housed in an epoxy resin case for stability and minimum weight, these coils are designed to meet all environmental requirements of MIL-C-15305A, Grade 1, Class B, thus yielding reliable operation over a temperature range of −65°C to +125°C. All units employ rugged construction techniques to insure that they will adequately meet military requirements.

For further information write to John Pfingsten at the firm.

Ultraviolet Spectrophotometer

A new descriptive bulletin on the DB ultraviolet spectrophotometer is available from the **Scientific and Process Instrument Div., Beckman Instruments, Inc.,** 2500 Fullerton Rd., Fullerton, Calif.

The 16-page brochure describes major features, outlines specifications, and explains the optical system and operating principles of the double-beam instrument. Numerous curves illustrating the DB's capabilities also are included.

(Continued on page 154A)

***A new slant** on making high accuracy DC measurements conveniently and economically!*

The model PC

PocketPot*

(PHOTO ACTUAL SIZE)

SENSITIVE RESEARCH'S NEW MINIATURE DC POTENTIOMETER!

The Model PC "PocketPot," is a stable, completely self contained, high accuracy DC potentiometer, with internal galvanometer, reference standard source, and direct "in line" readout. It is ideally suited for use as an infinite impedance calibrator or measuring instrument. Additional voltage and current ranges can be obtained by using it in combination with the Model PC-S, a switch controlled, .05% accurate "plug in" unit of the same size. Both models, when used together, may be easily held and operated in the palm of the hand!*

PocketPot*

SPECIFICATIONS

ACCURACY: ± .05% of reading or ± .5 mv., whichever is greater.

RANGES: 0—5.100 v. When used with PC-S 0—500 v. and 0—1 amp.

SENSITIVITY: Infinite resistance at null. When used with PC-S, 2,000 Ω/v.

RESOLUTION: Continuous, 1 mv. divisions on slide wire.

BATTERY OPERATED: Does not require an external power source.

READOUT: Direct "in line."

SIZE: 9" x 4¼" x 1¾". **WT:** 3 lbs.

PRICE: Model PC, $450.00; Model PC-S $175.00. F.O.B., New Rochelle, N.Y.

PocketPot*

Write for additional information or contact your nearest SRIC sales representative.

*Copyright U.S.A. Patent Applied for.

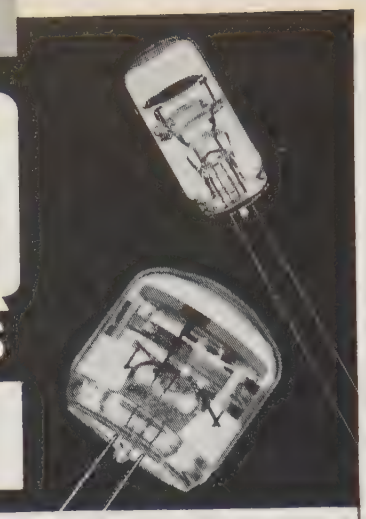
SENSITIVE RESEARCH
INSTRUMENT CORPORATION
NEW ROCHELLE, N. Y.

Symbol of Quality

ELECTRICAL INSTRUMENTS OF PRECISION SINCE 1927

EXCEPTIONALLY
HIGH Q
QUARTZ CRYSTALS

This group of High Q quartz crystals illustrates Bliley capability and quality. Each type meets extraordinary stability requirements in actual service . . . from time base reference to satellite equipment.



SERIES	BG93A	BG11AH-5	BG61AH	BG61AH	BG61AH-5
Freq.	1 mc	2.5 mc	3 mc	5 mc	5 mc
Q	1.5 x 10^6	4.5 x 10^6	.5 x 10^6	.5 x 10^6	2.5 x 10^6

These unusual characteristics are achieved with gold-plated, optically polished crystals, vacuum mounted in glass, to insure minimum long term aging.
Additional details are given in Bulletin 527 . . . sent on request.

Bliley . . . if you buy quality

BLILEY ELECTRIC COMPANY
UNION STATION BUILDING • ERIE, PENNSYLVANIA

HIGH VOLTAGE QUALITY RELIABILITY

POWER SUPPLIES

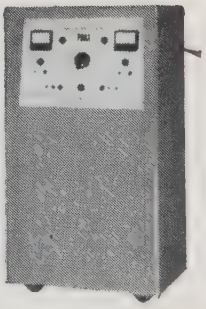
Standard power supplies and transformers available from stock. Others built to your specific requirements from stock components.

WRITE FOR COMPLETE SPECIFICATIONS

120 KV DC AT 5 MA

- Zero Start
- Reversible Polarity
- Input 115V-60 cycle
- Safety Interlock
- Electronic overload circuits provide full range protection
- Automatic stored energy discharge switch
- No exposed high voltage components

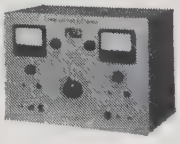
Completely instrumented supplies available from 1 KV to 300 KV DC and up to 50 KW.



Model PSC 120-5-2

30 KV DC AT 5 MA
OR 35 KV DC AT 1 MA

- Completely Self-Contained—Portable
- Ideal as Insulation Test Set and "Hi-Pot" TESTER
- No Exposed High Voltage Components
- Electronic overload circuits settable from 0-120% of rated current and voltage
- Units from 2.5 KV up to 60 KV available
- Many other features



Model PSC 30-5-1

DEL ELECTRONICS CORPORATION
521 HOMESTEAD AVENUE • MOUNT VERNON, N. Y. • OWENS 9-2000

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 152A)

Designed for precise qualitative and quantitative analyses in medical, industrial, and clinical applications, the DB may be used as a direct reading instrument or, with a recorder such as the Beckman Potentiometric Strip-Chart Recorder, as a ratio-recording spectrophotometer.

The instrument, offering both programmed and manual slit operation, has a wavelength range of 205–770 mμ and a resolving capacity of better than 5 angstroms.

For further information write to the Technical Information Department for Bulletin 779-B.

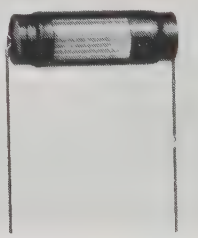
FXR Special Products Brochure

Special products of the **Microwave Division, FXR, Inc.,** 25–26 50th St., Woodside 77, N. Y. are displayed in a 4-page 2-color brochure. Among the special products illustrated and described in capsule form are a radar test set for missiles, high power rotary joint, stripline power divider, resistive loop directional coupler, three port switch, optical waveguide arc detector, high resolution frequency meter, power peaker, electrically variable attenuator, audio function generator, and others.

The brochure describes the multiple capabilities of the microwave division in special products design and also lists the engineering and production facilities of FXR for prototype or production runs. A copy of the brochure is available from the firm, or from FXR's various sales offices throughout the country.

Cadmium Sulphide Cell

National Semiconductors Limited, 230 Authier St., Montreal 9, Canada, announces the introduction of a new Cadmium Sulphide Photoconductive cell, NSL-33, for photometric applications, particularly where low light levels and low voltages are available.



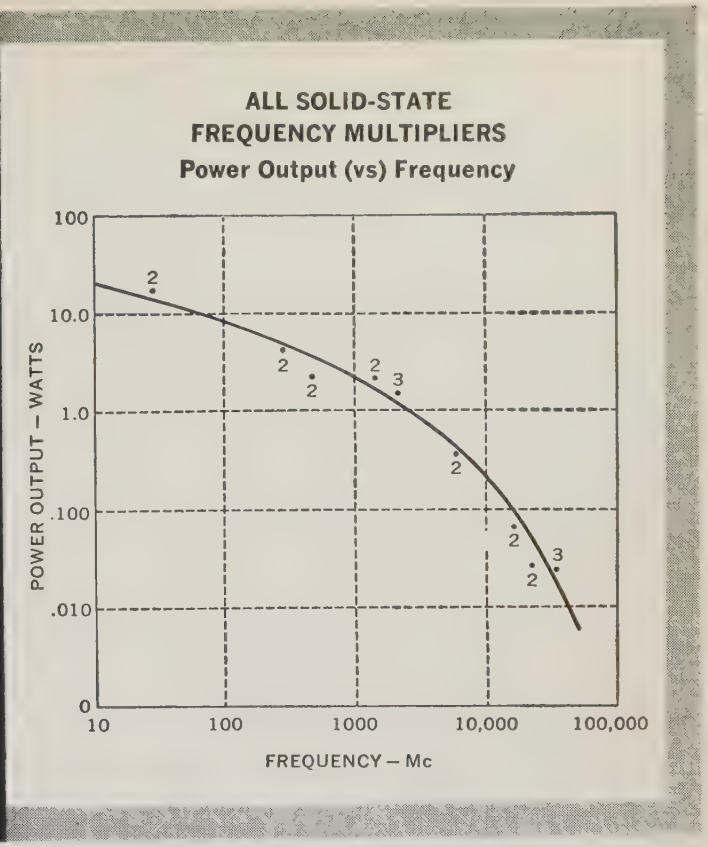
Spectral response resembles that of the human eye. Recommended illumination range, 0.01 to 1000 foot candles.

Tubular, hermetically sealed construction, 0.378″ diameter, 1.25″ long. Typical operating circuit involves a 1.34 volt battery and a meter movement.

(Continued on page 156A)

MICROWAVE ASSOCIATES PROGRESS REPORT

Microwave Power from Varactor Diodes



Efficient conversion of microwave power has been accomplished with a variety of new varactor frequency multipliers developed at the Waveguide Systems Division of Microwave Associates, Inc.

We have produced microwave power of several watts at UHF frequencies, several hundred milliwatts at X-band frequencies, and tens of milliwatts at Ka-band frequencies. The curve above indicates more accurately the power levels achieved by these Microwave Associates units. They employed doublers and triplers.

Efficiencies of these units range from 80 — 90% in the UHF region and from 20 — 30% at X-band. At present, the highest efficiencies are achieved at relatively narrow bandwidths (1%-2%). However, our capabilities are rapidly improving efficiencies for broader band operation. An example of a fixed-tuned broadband unit is a "tripler" which provides an output of 10 milliwatts over a 14% range at X-band.

Because of their efficiency and simplicity, these frequency multipliers are of considerable interest to systems engineers designing radar exciter circuits, low-power transmitters, stable local oscillator and paramp pump sources, and other circuits which require high frequency stability and exceptionally long life. These varactor multiplier circuits are generally passive, requiring neither tuning nor external bias voltage.

Our progress in producing efficient microwave power with all-solid-state techniques is related to performance of the most advanced high-power epitaxial varactors with significantly lower losses. The capabilities of Microwave Associates' Semiconductor Division in producing such varactors is a most positive asset. As this article is being printed, the multiplier performances shown here have already been exceeded.

We are also developing chains of these frequency multipliers to provide moderate amounts of power when driven by transistor oscillators. Efficiencies of these multiplier chains (RF output/DC input) are as good or better than equivalent klystron sources. Compactness and all-solid-state reliability are equally important benefits.

If you have an application for efficient varactor frequency multiplication or would like to discuss the very latest capabilities of these units, please write to Mr. Herbert Cox, Waveguide Systems Division. We'll be pleased to send you a new article on Varactor Frequency Multiplication by Mr. M. E. Hines.

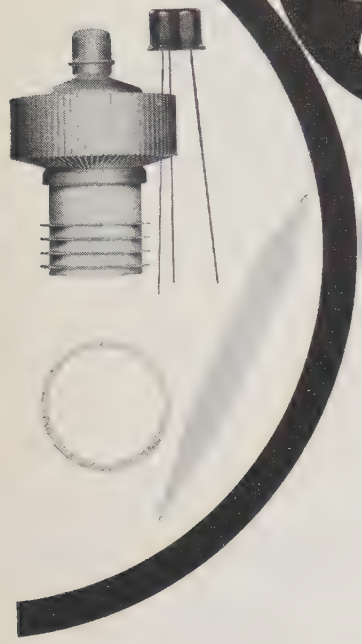
MICROWAVE ASSOCIATES, INC.
WAVEGUIDE SYSTEMS DIVISION

Burlington, Massachusetts • BRowning 2-3000
Western Union Fax • TWX: Burlington, Mass. 942

Export Sales: MICROWAVE INTERNATIONAL CORP.
36 W. 44th Street, N.Y.C., N.Y., U.S.A., Cable MICROKEN

Wesgo

Skilled specialists in premium ceramics for electronics!



WESGO delivers all the advantages of premium alumina ceramics

with the plus quality you need for vacuum tube envelopes and internal spacers, rf windows, heat sinks, high-voltage insulators, semiconductor components and a wide array of precision electronic applications.

Wesgo's specialization in ceramics for electronics gives you unique reliability in components you can depend on.

Three dense, vacuum-tight Wesgo alumina bodies, with up to 99.5% Al_2O_3, are strong, hard and abrasion resistant. They offer exceptional chemical inertness, high thermal conductivity and superior electrical properties, even at extremely high temperatures.

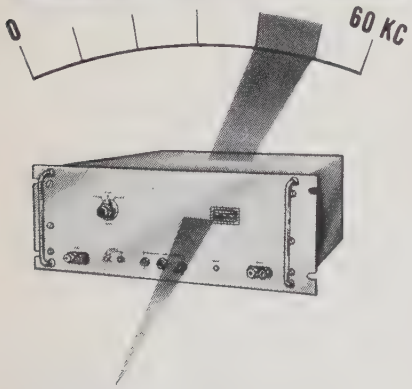
All Wesgo high alumina ceramics feature very low loss factor, high dielectric strength and high resistivity. All are available in sizes and shapes to meet your individual specifications. Manufacturing is to tight dimensional tolerances; ceramics are of uniform density, free from internal and surface defects. All are quality-controlled to meet unparalleled performance standards, with extraordinary care taken in each step of the manufacturing process.

Also available from Wesgo, a complete line of low vapor pressure, high purity, precious metal brazing alloys. Write today for information.

WESTERN GOLD & PLATINUM COMPANY

Dept. P-10, 525 Harbor Blvd. • Belmont, California • LYtell 3-3121

SPECTRUM TRANSLATOR



600 CPS to 18 KC

Variable positioning and/or inversion of audio spectrums for use with spectrum analyzer systems. Accommodates the translation of an input audio frequency baseband of zero to 60kc through an output spectrum window from 600 cps to 10kc with variable frequency positioning and selective channel inversion of the input signal. Provides spectrum processing for real time spectrum analysis equipment or other applications. Write for complete information.

RIXON

ELECTRONICS, INC.

2121 Industrial Parkway

Montgomery Industrial Park

Silver Spring, Md. 622-2121

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 154A)

Physical Sciences Booklet

A new fully-illustrated, 30-page booklet presents a comprehensive discussion of General Mills' extensive research in the physical sciences—research which has been concentrated in areas that are of fundamental interest. Electro-hydrodynamics, Millimicron Particle Technology, Sputtering in Plasmas, Upper Atmosphere Phenomena, Grain Boundaries and Dislocations, and Electron Mirror Microscopy—these and other activities are discussed in some detail in the booklet.

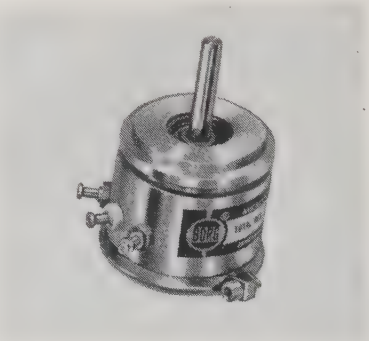
The various sections of the booklet deal with the scientific fields listed above; those in which the research scientists participate as they provide technical support for General Mills' electronic and electromechanical engineering and manufacturing activities, conduct company-sponsored basic research, and perform contract work for government and industry.

Copies of the booklet can be obtained by writing to: Advertising Manager, **General Mills, Inc.**, 1620 Central Ave., Minneapolis 13, Minn.

Precision Potentiometer

Borg Equipment Div., Amphenol-Borg Electronics Corp., 120 S. Main St., Janesville, Wis., announces a new series of $\frac{7}{8}$" diameter, single-turn precision potentiometers: the 2440 Series Micropot.

This new Micropot is linear over specified resistance ranges of from 10 to 50,000 ohms. Resolution ranges from 229 to 900 turns.



Housings and front bearing mounts are one-piece machined aluminum for maximum heat dissipation and permanent concentricity. All resistance elements are certified resistance alloys on thermosetting plastic cards. Contact and collector surfaces are precious metals specially selected to minimize noise, contact resistance and thermal effects.

Turret-type terminals are gold plated for solderability and corrosion-free shelf

(Continued on page 158A)

Max. MegW/cu. in.

ITT CERAMIC HYDROGEN THYRATRONS AND DIODES FOR MAXIMUM POWER/SIZE RATIO

No other hydrogen thyratron and diode line offers this wide choice of high-power types in the smallest possible tube envelopes. Made by ITT Kuthe Laboratories, most experienced maker of hydrogen-filled tubes, these ceramic thyratrons and diodes are the most complete line on the market today, with each type immediately available from production. The high-power thyratrons are designed to operate at high repetition rates and high temperatures; the high-power diodes may be used as hold-off diodes, inverse clippers and backswing clippers. Also available is an equally comprehensive line of glass thyratrons. The entire ITT hydrogen thyratron and diode program is aimed at the most demanding applications, particularly where long life and ruggedness are mandatory.

THYRATRONS	KU 70	KU 71	KU 72	KU 73	KU 74	KU 274
Performance Factor (x 10^9)	1.0	4.0	7.0	20.0	40.0	45.0
Peak Power Output, Megawatts	0.2	1.0	3.50	12.50	33.0	50.0
Forward Anode Voltage, Kilovolts	6.0	10.0	20.0	25.0	33.0	50.0
Peak Anode Current, Amperes	90	200	350	1000	2000	2000
Average Anode Current, Amperes	.100	.200	.300	1.5	4.0	4.0
Height, Inches	1.70	2.25	3.00	5.75	11.0	12.0
Diameter, Inches	1.15	1.37	1.75	3.00	4.50	4.50

DIODES	KU 92	KU 93	KU 94
Peak Inverse Voltage, Clipper, Kilovolts	25.0	33.0	33.0
Peak Forward Current, Clipper, Amperes	300	500	2000
Max. Average Current, Clipper, Amperes	.200	.500	2.0
Peak Inverse Voltage, Rectifier, Kilovolts	15.0	20.0	20.0
Peak Forward Current, Rectifier, Amperes	2.0	8.0	16.0
Max. Average Current, Rectifier, Amperes	.50	2.0	4.0
Height, Inches	3.00	6.00	8.25
Diameter, Inches	1.75	3.00	4.50

Write for information on the complete line of ITT hydrogen thyratrons and diodes. Application assistance is available for your specific requirements.

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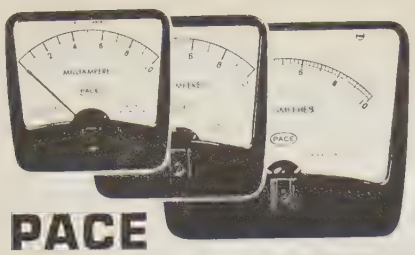

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NEWS New Products

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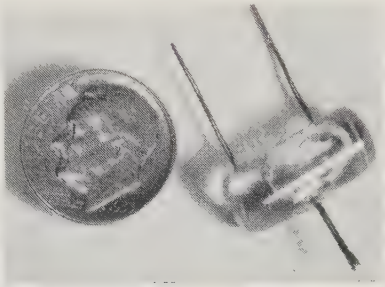
(Continued from page 156A)

life. All 2440 Series Micropot models meet a full range of military specifications including vibration, acceleration, shock, sand and dust, temperature cycling, humidity, fungus, salt spray, and so forth.

Maximum power rating is 3 watts. Temperature range is from −55°C to +125° C. Gangable up to six units on a common shaft, the potentiometer is offered in either servo or bushing mounts.

Vertical Sensing Element



Kearfott Div., General Precision, Inc., 1150 McBride Ave., Little Falls, N. J., has announced the production of the 228571-1 Vertical Sensing Element, a single axis, snap action, bubble type, electrolytic unit which can drive torque motors directly without amplifiers. An "on-off" snap action device, this instrument cannot be nulled, and it has a tilt angle of 0.75° (+30 arc minutes spread from full scale in one direction to full scale in the opposite direction), 18 volts differential output voltage, and 18 ma maximum current through either contact. This unit has no null and operates between 75 minutes and 135 minutes of arc. This is one of a series of subminiature vertical sensing elements produced by Kearfott's Glass Lab., a facility which combines glass design techniques with production capability to provide gyrodynamic components.

Low-Level Beta Counting System

A combination of four detector units capable of simultaneous operation is obtainable in a transistorized low-level beta counting system announced by **Transnuclear Corp.,** 120 Woodland Ave., Westwood, N. J. The system was developed by Isotopes, Inc., and tested for more than two years in the analysis of thousands of samples generated by the AEC's fallout measuring program where low backgrounds, stability and reproducibility were of prime importance.

The system also has interchangeable plug-in units, unitized counting assemblies and ¼ count per minute for 1″ diameter high sensitivity detectors. Each unit has a sample holder, sample detector, guard detector and scaler. A user has the option of installing 1, 2, 3 or 4 units with expansion made possible (up to four) by purchase of extra components and simple field conversion without return to the manufacturer.

Depending on the number of detector/scaler units, the system ranges in cost from $5800 to $7750. Alpha detectors with preamplifiers are also available as accessories. Specification Folder CLL/1-4 is available from Nuclear Industries, 10 Holland Court, Valley Stream, N. Y., exclusive distributor of the product.

Electronic Generator

Industrial Test Equipment Co., 55 E. 11th St., New York 3, N. Y., has expanded its line of Powertron electronic generators to include variable frequency 3-phase units with output powers ranging from 10 va thru 3000 va. Typical is the Model 150-3 (illustrated) which has a frequency range of 45 cps to 2000 cps and a total output power of 480 va. The frequency range is covered on two (2) range scales which pro-

(Continued on page 160A)

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RESOLUTION
REAL TIME
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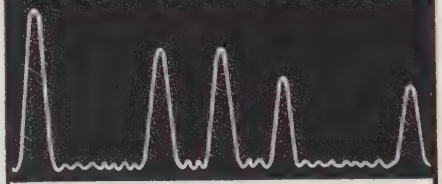

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This Very Low Frequency Spectrum Analyzer provides up to 1000 line resolution for frequency ranges from 0-1 cps to 0-1000 cps. Analysis time in all frequency ranges is 1 second. Transistorized Models SA-12 and SA-14 will provide expanded frequency range and with GASL auxiliary equipment offer power density and normal density capability.

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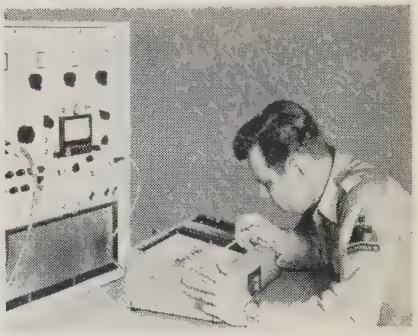
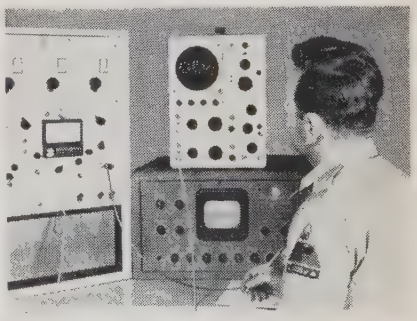


RADIO INTERFERENCE — FIELD INTENSITY MEASURING EQUIPMENT



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  ELIMINATES need for manual tuning
- **AUTOMATIC SPECTRUM SIGNATURES** — with use of X-Y (Analog) Output and AUTOSCAN
  Operator can evaluate X-Y plot while NM-62A continues its data collection

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Sensitivity:	2 to 4 microvolts with 500 kc bandwidth; 6 to 12 microvolts with 5 mc bandwidth
Bandwidths:	500 kc; 5 mc
VSWR:	1.5 to 1 with 50 ohms input impedance
Oscillator Radiation:	18 micromicrowatts
Case Shielding:	90 db

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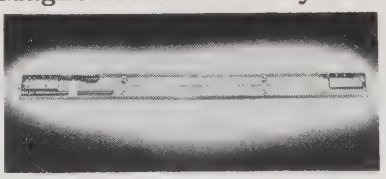
(Continued from page 158A)

vide high resolution and simple setting. New phase shifting techniques provide accurate 3-phase displacement of the output over the entire frequency range. Output distortion and regulation are less than 1%. All three output phases are circuit breaker protected against overload. Units are also available with standard 350–450 cps frequency range.

Magnetostrictive Delay Line



Designed by **ESC Electronics Corp.**, 534 Bergen Blvd., Palisades Park, N. J., for applications requiring higher delay time as in computers, data processing and airborne instruments, the model 20Mol magnetostrictive delay line provides a time delay of 60 μsec. Adjustment range of ±3 μsec; input impedance of 550 ohms and output impedance of 2000 ohms; attenuation of 50 db and a signal-to-noise ratio of 12:1 are also features of this new model.

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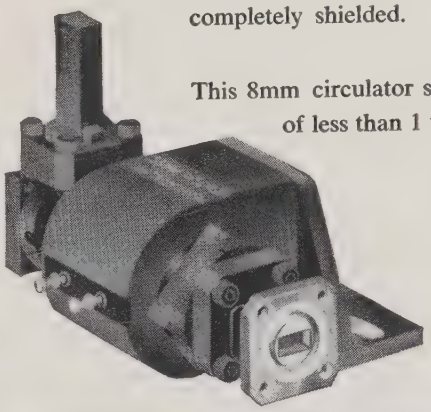

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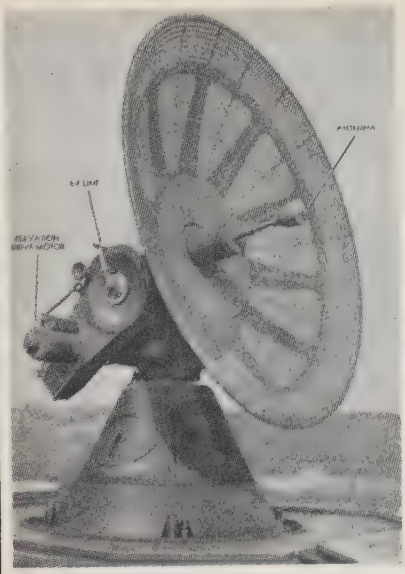

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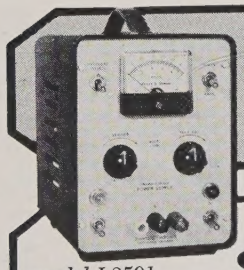
DC POWER SUPPLY



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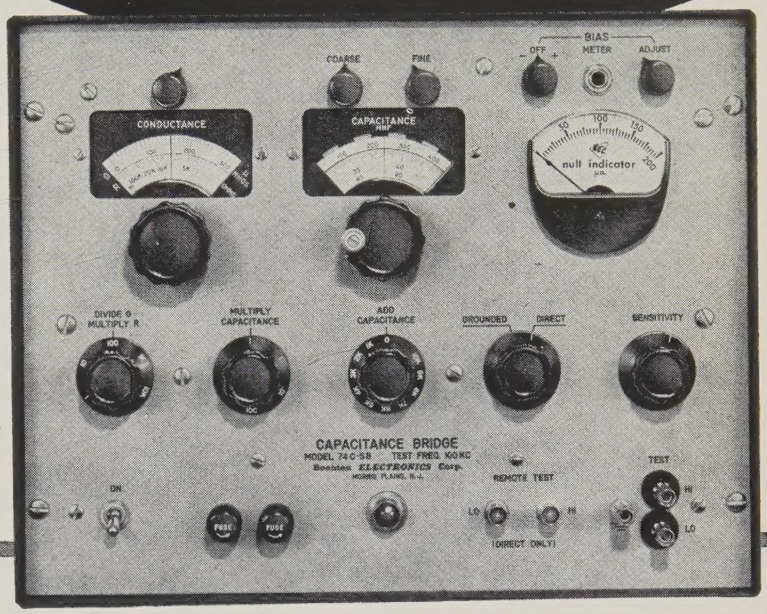

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(INSENSITIVE TO GROUNDED CAPACITANCE)
CAPACITANCE BRIDGES

MODEL 74C
- 100 KC Test Frequency
- 0.0002-11,000 μμf Generally 0.25%
- 1000 ohms to 1000 megohms Shunt Resistance
- 0.001 to 1000 μmhos Conductance

Price $935

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MODEL 74C-S8 (Shown)
- With −5 to +100V DC Bias for Diode Testing

Price $995

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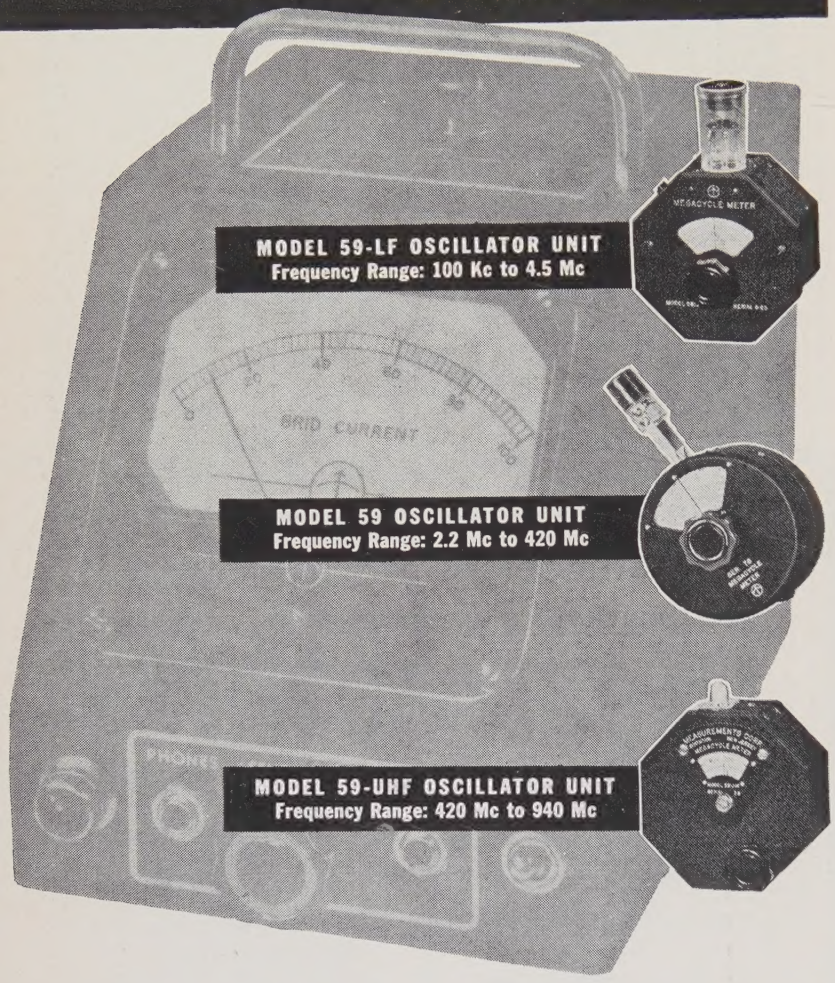
MEASUREMENTS' MEGACYCLE METER

0.1 Mc to 940.0 Mc


MODEL 59-LF OSCILLATOR UNIT
Frequency Range: 100 Kc to 4.5 Mc
MODEL 59 OSCILLATOR UNIT
Frequency Range: 2.2 Mc to 420 Mc
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Frequency Range: 420 Mc to 940 Mc
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MEASUREMENTS' MEGACYCLE METER is now available in your choice of three oscillator heads, covering a total frequency range of 100 Kc to 940 Mc. Specific range of each oscillator unit is indicated in the illustrations shown. The Power Supply is a compact, light-weight, portable unit, comprising a full-wave rectifier with voltage regulator tube.

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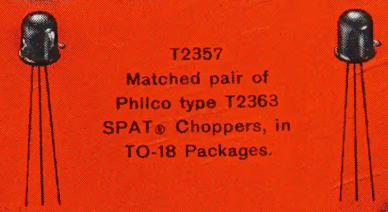
NEW PHILCO SILICON CHOPPERS



T2357
Matched pair of
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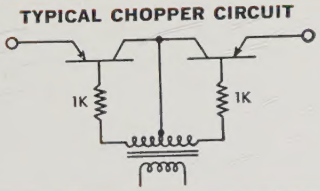
SPAT® Choppers, in
TO-18 Packages.

With SPAT* Matched-Pair Uniformity Bring High Fidelity To Low Level Switching!

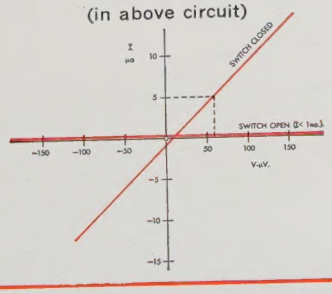
T2363 CHARACTERISTICS

Emitter Voltage, BV_{ECO}	−30 volts
Collector Cutoff Current $I_{CBO}(V_{CB} = -10V)$	0.001 μa max.
Emitter Collector Current $I_{EBO}(V_{EB} = -10V)$	0.001 μa max.
Offset Voltage $V_{EC}(I_B = -200\ \mu a, I_E = 0)$	1.5 mv max.
Offset Voltage V_{EC}(T2357 Matched Pair, $I_B = -1ma$ at all temperatures from 25° to 85° C)	50 μv max.

TYPICAL CHOPPER CIRCUIT



T2363: I-V CHARACTERISTIC
(in above circuit)



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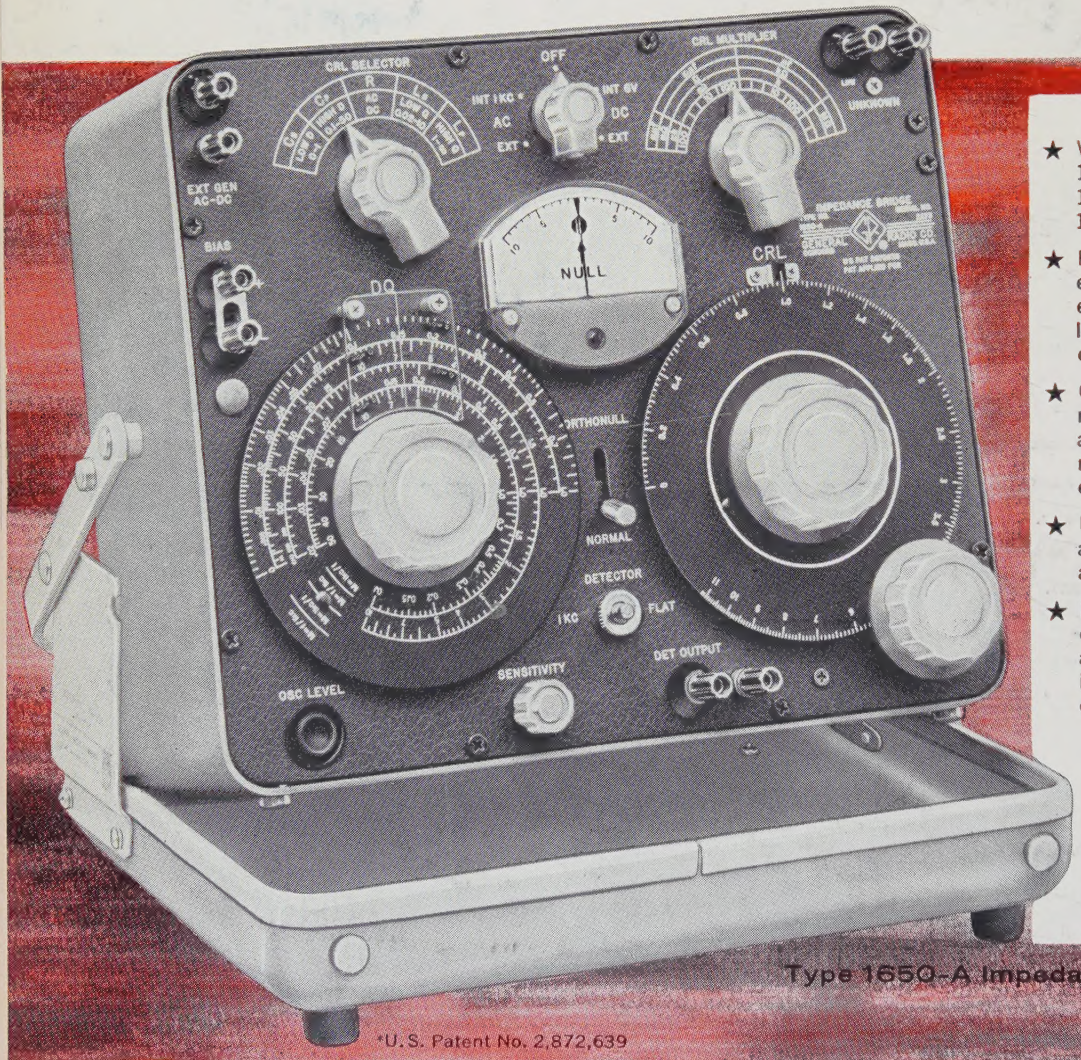
A FUNDAMENTAL INSTRUMENT

... useful for measurements ranging from interelectrode capacitance of semiconductors to impedance of transformers weighing several tons.



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